

MODELING, SIMULATION, AND DESIGN OF
HIGH-SPEED HETERODYNE/QUANTUM NONLINEAR TRANSISTORS
FOR HIGH-PERFORMANCE INTEGRATED CIRCUITS

By

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*This dissertation is dedicated to my wife and friend, Ayun,
and to my children, Albert and Benjamin.*

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**MODELING, SIMULATION, AND DESIGN OF
SILICON-BASED HETEROJUNCTION BIPOLAR TRANSISTORS
FOR HIGH PERFORMANCE INTEGRATED CIRCUITS**

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This dissertation presents a physical but compact semi-numerical charge-based SiGe-base heterojunction bipolar transistor (HBT) model, which is also applicable to advanced Si bipolar junction transistors (BJTs) in which bandgap narrowing can occur due to heavy base doping. The base-emitter transport analysis gives a characterization of the transport current for all injection levels with a unified accounting for bandgap-narrowing due to Ge and/or heavy doping. The collector transport analysis, which accounts for all possible quasi-collector charge conditions including quasi-inversion and potential barrier formation due to the base-collector heterojunction, is physically linked to the base analysis via proper moving boundary conditions affected by carriers (ions or electrons) recombination. Transport currents and quasi-static stored charges (and charging currents) are characterized in terms of technological parameters, which relate directly to device structure. The resulting model is implemented in the circuit simulator SPICE3-H to create an HBT version of MINIPICE, a semi-numerical mixed mode simulator. Non-local accounting for impact-ionization current generated in the collector region, based on analytic solution of a simplified energy balance equation involving the electric field

distribution provided by the collector analysis, is implemented as post-processing in the model routine. The new HBT model is verified by application to an aggressive SiGe-base HBT technology. This application reveals heretofore unacknowledged degrading effects on EFTHBT speed performance due to current-induced perturbation of the collector-base junction space charge region that arise prior to base pushout. Inverse base width modulation (IBWM) – a widening of the quasi-neutral base width – and collector space-charge-region widening (SCRW) in highly scaled SiGe and SiGeC are shown, by simulations and measurements, to be important mechanisms limiting device speed at high currents. Utility of the SiGe-base HBT modeling MEMFICE for technology CAD is demonstrated by using the tool in a device design optimization exercise for both digital and analog integrated circuit applications.

CHAPTER 1 INTRODUCTION

Recent impressive progress in SiGe base heterojunction bipolar transistor (HBT) technology has significantly extended the device performance limit imposed by the conventional homojunction bipolar junction transistor (BJT) [Mar03, Bur03]. The numerous options for designing high-performance devices provided by the SiGe alloy heterojunction on the silicon substrate, combined with the significant reduction of parasitics afforded by the self-aligned double polysilicon process, enabled the record high unity current gain frequency (f_T) of 115 GHz, and BCE propagation delay well below 10 ps [Kus03b]. In addition to the design flexibility allowed by the bandgap engineering in the base region, the process compatibility with the existing well established silicon based integrated-circuit (IC) technology is another key advantage. Recent progress in ultra high-voltage channel-length dependent (UHVCLD) and atmospheric-pressure GCD (APGCD) technologies leading to the mass production of SiGe-based HBT ICs has encouraged VLSI chip production on 200mm-diameter silicon wafers [Mar03, Bur03].

In contrast to the rapid advance in SiGe base HBT processing technology, there has been little work on device model development for circuit simulation, which is vital for design optimization of the device and circuits, as well as of the technology. Most of the work related to modeling for this technology has focused on the implementation of the bandgap narrowing due to the self-manual device simulation, and a few crude compact models based on the

General-Purpose BJT model have been suggested in terms of no-average bandgap narrowing across the bias regime. The General-Purpose model, which has been the work horse for almost BJT circuit simulation for decades, is so simplified that it does require extensive parameter extraction based on measured data, and still is not reliable in all cases, for example, where high-injection effects are prevalent in the BJT. Further, the increasing demand for predictive circuit (and mixed-mode device/circuit) simulation in IC technology CAD (PCAD) systems is not met with the General-Purpose model.

A viable approach to efficient and reliable mixed-mode device/circuit simulation for bipolar IC/PCAD was recently suggested. It involved developing a physical, semi-empirical, charge-based (transportation) BJT model based on technological (structure-dependent) parameters and implementing it into SPICE2.4 source code, creating Mixed-Mode DEVICE (MMSPICE) [Joshi]. The BJT model algorithm is based on regional analysis of ambipolar transport, subject to moving boundary conditions, and their linking to semi-empirical solve the system of carrier-transport equations. The algorithm thus must have ensured continuity of all currents and charges through the entire range of operating modes, including quasi-saturation. Despite the success of early versions of MMSPICE in simulating the quasi-saturation effects in the BJT in a unified way they had many drawbacks related to current and charge discontinuities across the model boundaries yielding abnormality in terminal current-voltage characteristics [Hosoi]. Further, these convergence problems as well as the computational inefficiency in the quasi-saturation modes severely limited the model utility in high level IC simulations. In addition to these computational drawbacks, the model neglected the bandgap narrowing due to heavy doping in the base region, an effect which is indeed significant in advanced scaled Bi BJTs [Greene].

The research discussed in this dissertation effectively removes the drawbacks of the MMSPICE-1.1T model, and further extends the model formulation to provide a unified characterization of the advanced SiC NTE, accounting for the bandgap narrowing in the base due to heavy-doping, models the SiGe-base EBT, accounting for the bandgap narrowing due to Ge as well as heavy doping. The resulting enhanced version of MMSPICE is shown to be a computationally efficient mixed-mode device-level simulation tool for SiGe-base EBT IC TCAD.

In Chapter 3, with an objective of first enhancing MMSPICE to characterize its computational drawbacks and to make it the basis for the SiGe-base EBT model, we survey general approaches and techniques for optimizing regional semi-numerical device models with regard to accuracy, convergence, and computational efficiency. Emphasis is placed on refining the physical terms underlying the discontinuity of currents and charges. The details of the model-region modification implemented in MMSPICE-1.8 [Kra04] are presented, and the resulting computational improvements in the tool are demonstrated.

In Chapter 4, we develop the physical charge-based SiGe EBT model for circuit simulation that properly accounts for the nonuniform energy gap in the base, as well as the nonuniform doping density, both of which influence the electron transit time and hence the device speed. The model is applicable to both single- and double-heterojunction transistors. As the basis for the model development, we use the EBT model in MMSPICE-1.8, which is reviewed. The physical hot-carrier model for the SiGe-base EBT is developed by extending the present underlying work [Kra03, Kra05, Jacobs]. The new base analysis assumes an exponential doping profile in the quasi-neutral base (QNB) and a linear bandgap variation, both of which are representative of an actual device

In fact, bandgap narrowing in highly doped silicon tends to vary logarithmically with respect to the doping density, the analysis enables validation of the HBT model, with lower Ge distribution, and that for the scaled Si HBT with significant bandgap narrowing at the base. Assuming small hole current in the *npn*-device, we describe the effective field acting on the electrons. Then integrating the electron current equation across the quasi-neutral base region for both cases of low and high injection, we express the collector current in closed form terms of the boundary conditions at the moving emitter and collector sides of the QNBL. Unlike the Gummel-Poon HBT model for which empirical transit times are needed to relate stored charge to current, the analysis also enables integration of electron density over the QNBL to characterize the charge density.

The base model is physically linked to the analysis of the carrier transport in the *np*-collector region, which physically accounts for space saturation in *p*-₁ base widening). An extension of the MINIPICE-1.1 collector analysis to account for energy barrier formation [Choi9, Yoon0] in the conduction band at high currents due to bandgap discontinuity at the collector junction satisfies the boundary conditions linking the analysis to that of the base. The energy barrier formation in the conduction band, which causes excess carrier accumulation in the base and an increase of the base transit time leading to device speed degradation, has been previously accounted for in numerical device simulation [Yoon0] but not in any current simulator. We give a physical model for the barrier formation at each base point. The complete carrier transport analysis, which involves iterative solution, forms the basis for the complete physical, charge-based, large-signal double HBT (QDFT) model. The base current model and the regional charge modeling, with updated partitioning between the

emitter and collector terminals, are discussed as well in the presentation of the network representation of the model.

Accurate prediction of the built-in voltage BV_{BIO} of a bipolar technology becomes more important for both circuit and device design as the technology is scaled since the system power supply is not decreased to the extent that BV_{BIO} is reduced. Conventionally the impact ionization, which defines BV_{BIO} , is modeled with a local-field characterization [Bolt83] of the ionization rate, $\alpha(E)$, which is based on the presumption that carriers are instantaneously thermalized to the steady state kinetic energy corresponding to the local electric field intensity. In a spatially varying electric field, however, the carrier energy lags the field because of finite energy relaxation time [Bolt83]. This is the case for highly scaled BJT's in which very high electric fields and field gradients exist in the np -injection region. To properly model effects of impact ionization in advanced BJT's and their dependence on device structure, α must be characterized in terms of the (average) carrier (electron for the npn BJT) energy, or temperature T_e . Since T_e lags E , the local field (empirical) representation of α will overestimate the impact ionization and thus underestimates BV_{BIO} .

In Chapter 4, with an objective of predictive circuit simulation, we develop, and implement for the first time, a non-local characterization for impact-ionization current in our 3-D pnp model, applicable to advanced npn BJT's as well. We use a previously described simplification [Bolt83] of the energy-balance equation to define $T_e(x)$ from $E(x)$, and subsequently $\alpha(T_e)$. This analysis model has been successfully used in transient device simulation [Bolt83] in which the electric field distribution is part of the solution. However, it has not been properly incorporated in device models for circuit simulation since the conventional circuit models do not give a physical characterization of

$E(x)$. The averaging for $T_{\text{eff}}(\mu)$ in [Kus03] for bipolar current simulation is not generally valid because it uses $E(x)$ defined by the simple $p\text{-}n$ junction diode two approximation. The field distribution in an actual bipolar transistor is more complex due to prevalent high current-induced quasi-saturation and conductivity modulation in the collector [Jus07].

The charge-based BJT model in MMSPICE [Jus03, Jus03, Kus03] physically accounts for all possible charge conditions in the collector region for all bias conditions. We use it and the general $E(x)$ it provides as part of every line-point relation for implementing the neutral model for impact-ionization current. The implementation is done in post-processing in the model routine, and hence does not significantly increase the complexity of the BJT model in the circuit simulator. The new level model in MMSPICE is verified by applications to two advanced bipolar technologies, including an aggressive SiGe base BJT process [Ug00]. The utility of the tool in assessing bipolar design tradeoffs involving impact ionization is demonstrated by device/circuit simulations.

In Chapter 5, the model implementation, which includes writing the model algorithm into SPICE2.6 source code to create MMSPICE 3.0 and the development of new model parameter extraction formulas for SUMM [Kus03], is described. The MMSPICE-3.0 BJT model is then assessed by applications to the highly scaled NTE SiGe-base BJT technology [Ug00] mentioned above. The (measured) minimum region doping profile of this technology is used to evaluate model parameters, without any optimization, using SUMM. Results show favorable comparisons of measured and model-simulated $G_{\text{sat}}(\mu)$ plots, $I_C(V_C)$ and $I_{\text{BCE}}(V_C)$ characteristics, with and without the Ge incorporation, and hence verify that our model reliably accounts for the bandgap-narrowing effect due to the Ge as well as the heavy doping in the base region.

Through the measured and simulated α characterizations, we found that the peak current where the maximum device speed occurs is significantly lower than the theoretically-cast for the base product. Degrading effects on BJT speed performance due to parasitic-induced perturbation of the collector-base junction space-charge region (SCR) prior to base product are identified, analyzed, and assessed. Drives base-width modulation (EBWM) – a widening of the space-extended base width – and collector SCR-width widening (SCSW) on highly scaled BJTs and HBTs are identified as important mechanisms governing device speed-degradation at high currents. EBWM, which increases the base transit time, is described analytically to distinguish it from base product, or quasi-saturation. NOISEFICE simulations of the mentioned SiGe-base HBT technology, supported by measured data, show that (the speed- f_T) degradation associated with EBWM is significant at currents well below the onset of base product, which underlies the speed-degradation of lower scaled BJTs. Simulations of further scaled devices show that SCSW can be the predominant mechanism of speed-degradation at high current densities.

In Chapter 6, with the objective of demonstrating the utility of NOISEFICE-1.0 as a device as well as circuit design-aid tool, we use our HBT model to suggest optimal device design for high-performance digital and analog IC applications. Various design considerations for optimal HBT structures are discussed focusing on the effect of the grading on device speed and Early voltage. We start the simulation exercise with references to the HBT technology, and vary the base and collector structural parameters, which are accessible in our model. We compare the optimized devices with representative published data to support the model validity as a predictive technology design-aid tool, and show circuit simulation results reflecting highly superior figures of merit for typical digital and analog ICs.

In Appendix A, a brief description of the effective fields in the base region, which underlies the HBT base analysis (Chapter 3), is presented with appropriate assumptions adopted. In Appendix B, the electron temperature formulas for the implementation of the non-local accounting for recombination current (Chapter 4) in the model routine are provided. Finally in Appendix C, physically motivated expressions pertaining to the design of SiGe HBTs with uniform up-base doping and lower Ge grading (Chapter 4) are derived.

CHAPTER 2 ENHANCEMENT OF MINIPICE BJT MODEL

2.1 Introduction

The novel concept of semi-numerical mixed mode formulation simulation was recently presented through the development of MINIPICE (Jantha) for advanced bipolar technology CAD. MINIPICE was created by developing a physical, semi-numerical, charge-based BJT model having technological (structure-dependent) parameters; by defining a model algorithm to numerically (via Newton-like iterative method) solve the system of model equations; and by implementing the model in SPICE2 source code. The BJT model is based on regional analysis of ambipolar transport, subject to moving boundary conditions. The algorithm then must ensure continuity of all currents and charges throughout the entire range of operating modes (α , for all V_{BE} and V_{BC}), for all possible regional boundary conditions, as well as define the current control for the numerical solution.

Although the previous version of the simulator, MINIPICE 1.2, simulates successfully the quasi-static-state effects in the BJT in a unified way, it has shown several drawbacks in its performance causing problems in actual applications. A notable example is that the predicted current-voltage character of a BJT's α , the $I_C - V_{CE}$ curve with I_B constant, has some unrealistic features for some range of V_{CE} in the high current region of operation. The flat characteristic implies that the derivative of the current α , output conductance, has a discontinuity, which could limit the utility of MINIPICE in analog circuit

simulation. For steady CAC, the small signal equivalent-circuit models are strongly dependent on the voltage derivatives of the terminal characteristics, and hence these derivatives as well as the terminal currents must be everywhere continuous.

Furthermore, MMSPICE-1.1 has shown convergence problems, which portend discontinuities in the predicted IFT currents and charges. Also, the model tends to be computationally inefficient, especially when the transistor is in quasi-saturation or saturation. These drawbacks seem to arise mainly from inconsistent application of the underlying model assumptions and from the non-physical definition of some parameters in the model equations.

With an objective of enhancing MMSPICE to eliminate its drawbacks, the work described in this chapter covers general approaches and techniques for optimizing regional small-signal device models with regard to accuracy, convergence, and computational efficiency. It aims at effecting the aforementioned continuity in the model solution via use of strictly physical characteristics, avoiding the need for “smoothing functions” as in [16-18a]. The work involves modifications of the regional charge expressions, a new physical definition of the quasi-Fermi level (QFL) separation V_{qfl} at the transition region between the quasi-neutral (QNE) and carrier-injected space-charge (SC) regions in the *npn* collector [16-18a], definition of the electron carrier density $n(x_c)$ at the base-side edge of the collector junction space-charge region, and finally a substantial modification of the steady-state, which renders the simulator computationally more efficient in the saturation modes of IFT operation. The result is a re-written version of the model within QSPICE and a new release of the simulator, MMSPICE-1.2.

To put the new work in proper perspective, the key model equations implemented in MMSPICE-1.2 are first reviewed. Then the details of the

model-driven modifications implemented in MONPICE-1.3 are presented. Next, device and circuit simulation results are given to demonstrate the significant improvement in the simulator. Finally, the main accomplishments of this work are summarized.

2.2. Features of the MONPICE-1.3 BJT Model

The BJT model in MONPICE 1.3 [Jost03a] was developed with emphasis on the high-current quasi-saturation mode of operation. The high-current effects materialize predominantly in the epitaxial collector because of its relatively low doping concentration. These effects can be categorized under two physically different operational modes. One is the normal forward-active mode in which a finite collector-positive BCS is established; the other is the so-called quasi-saturation mode in which the BCS is collapsed and large push-out effects occur. The quasi-saturation mode can be further classified as short, pseudo-short, and non-short, depending on the degree of carrier velocity saturation [Jost03a].

The BJT model equations are implemented in QSPICE. The input modeling based on the one-dimensional device structure shown in Fig. 2-2, which represents the advanced epitaxial-collector device. The assumed electron drift velocity vs. electric field curve is illustrated in Fig. 2-3; E_0 is a critical field above which the velocity is saturated at v_0 . The velocity is proportional to the electric field when the field intensity is less than E_0 ; the proportionality constant is the mobility μ_{n0} . In the model formulation, the electron velocity is assumed to be saturated throughout any BCB, irrespective of $J(x)$. Hence the carrier density in the BCBs is approximated everywhere as $J_0/(qA v_0)$, where A is the (1-D) BJT area; carrier diffusion is neglected in the BCBs. As we shall

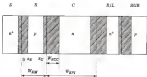


Fig. 3-1. One-dimensional structure of an $n^+p-p^+n^+$ junction bipolar transistor with its regional boundaries labeled and regional charges indicated.

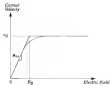


Fig. 3-8. Angeli model for carrier drift velocity-electric field dependence, referenced to the optical dependence.
 Note that $v_d = \mu_{eff}E$ for $E < E_0$.

ing, this leads for the SCB analysis under the same of the aforementioned problems with MOSFET-LE.

The MOSFET-LE model comprises two primary regional analyses and linkage of the results through defined conditions at the regional boundaries, which occur with variations in time. The base analysis gives a transport (collector) current and the associated regional changes for all injection levels. It assumes an exponential doping profile, and is based on the electron carrier densities at the moving quasi-neutral region boundaries. The collector analysis is more complicated, even though the eye doping is assumed uniform. If the junction SCB is found to exist, the collector analysis yields the position x_c (see Fig. 3-4) of the moving boundary and the electron-carrier density $n(x_c)$ for the base analysis; if no junction SCB is found to exist, i.e., if the LEF is in quasi-saturation (with a possible current-induced SCB in the eye collector), it yields the quasi-Fermi level separation at the metallurgical base-collector junction, via a modification of Kuhl's analysis [Kul88], and the electron-carrier density at the junction.

In determining the operational mode of the LEF, the key equation is the expression for the collector junction SCB width W_{JCC} (see Fig. 3-4). With the assumption that the integral of the electric field across the entire collector region equals the junction built-in potential ϕ_{jc} plus the reverse-bias voltage $-V_{BC}$, W_{JCC} can be expressed in terms of J_c by applying Poisson's equation to the junction SCB (see (10))

$$W_{JCC} = \left[\frac{2qA\phi_{jc}/A_{jc} - V_{BC} - J_c R_{JCC}}{I_B - I_C} \right]^2 \quad (10)$$

where ϵ is the permittivity of silicon, R_{BCE} is the base-collector bulk resistance given by $W_{BCE}/(\epsilon\epsilon_0 q N_{BCE})$, and I_C is the critical onset current given by $\epsilon\epsilon_0 q N_{BCE}$, where N_{BCE} is the base-collector doping concentration.

The presumed electric-field pattern for the derivation of (3-1) is shown in Fig. 3-3. The assumption that the injected carrier velocity in the BCL is everywhere saturated is used, and hence the effective space-charge is given by $N_{BCE} - I_C/(qV_{BC})$. Note that (3-1) is valid only for $I_C < I_C$ and $\phi_C > V_{BC}$. For $I_C > I_C$, a current-induced BCL forms away from the base-collector junction and modulates (3-1). For $\phi_C < V_{BC}$, the base-collector junction is heavily forward-biased and the entire ep collector is space-charged.

When $I_C < I_C$, Ψ_{BCE} via either expansion shrinks with increasing I_C due to the two counteracting terms involving I_C in (3-1); the term in the numerator accounts for charge drop in the quasi-neutral region, whereas the term in the denominator represents the effect of space-charge modulation by current-carrying electrons. The dominance of one effect over the other depends on the value of V_{BC} . A critical value of V_{BC} can be derived from the derivative of (3-1) with respect to I_C , and is given by $V_{CBET} = \phi_C - I_B R_{BCE}$. If $V_{BC} < V_{CBET}$, Ψ_{BCE} expands with increasing I_C , it contracts otherwise. From the above considerations, the condition at which the junction BCL collapses completely, as well as the condition at which the entire ep collector becomes space-charged can be defined. They are, respectively,

$$W_{BCE} = 0 \quad \text{for } I_C > I_{BCE} \text{ and } V_{CBET} < V_{BC}$$

and

$$W_{BCE} = W_{BCE} \quad \text{for } I_C > I_{BCE} \text{ and } V_{BC} < V_{CBET}$$



Fig. 2-3 Simple model of electric field across opt collector layer for the calculation of $\bar{W}_{sheath}$. Note that $E_{col} = E_0' / (n^2 \epsilon_0 \epsilon_{col} R_{eff})$.

where

$$I_{\text{SCC}} = I_0 - \frac{V_{\text{SC}} - V_{\text{CCST}}}{R_{\text{SCC}}} = \frac{q_p - V_{\text{SC}}}{R_{\text{SCC}}} \quad (3-2)$$

and

$$I_{\text{SCCmax}} = I_0 + \frac{1}{R_{\text{CC}} - R_{\text{SCC}}} (V_{\text{SC}} - V_{\text{CCST}}) \quad (3-3)$$

with $R_{\text{CC}} = R_{\text{SCC}}' (R_{\text{SC}}/r_d)$

With the positive SCB collapsed, a current-induced SCB can be formed away from the collector junction if $I_0 > I_{\text{SC}}$, a possible electric field pattern for this case is the top collector shown in Fig. 3-4. With the assumption that the induced SCB exists where $E(x) > E_{\text{SC}}$, and hence the majority electron velocity is saturated, that the electron current flows predominantly by drift, and that the minority hole density in the SCB is negligible compared to the electron density, the SCB width W_{SCB} can be expressed as (derived)

$$W_{\text{SCB}} = \frac{2q_p (V_0 - V_{\text{SC}})}{qE_0\rho_p + \sqrt{(qE_0\rho_p)^2 + \frac{2qV_0}{A} (I_0 - I_{\text{SC}}) (V_0 - V_{\text{SC}})}} \quad (3-4)$$

In deriving (3-4), the potential drop in the QSE is neglected. From this equation, it can be argued that W_{SCB} increases with decreasing I_0 for a given V_{SC} . Thus, the following criterion can be given.

$$W_{\text{SCB}} = W_{\text{SC}} \quad \text{for } I_0 < I_{\text{SCCmax}}$$

where

$$I_{\text{SCCmax}} = I_0 - \frac{V_{\text{SC}} - V_{\text{CCST}}}{R_{\text{CC}}} \quad (3-5)$$

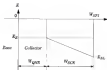


Fig. 3-4 Simple model of electric field across emitter-collector layer for the calculation of $\mathcal{W}_{ECC}$. Note that the electric field is assumed to be negligible within $\mathcal{W}_{ECC}$.

Note that this criterion is meaningful only for $V_{BC} < V_{BGR}$, for which $I_{BGRmax} > I_0$. For $V_{BC} > V_{BGR}$, the current-induced SCE forms when $I_C > I_0$. In any case for $I_C = I_0$, the current-induced SCE cannot be formed, and (2-4) is invalid.

Equations (2-1) through (2-5) are the key expressions used to define the modes of operation, which are distinguished by three charge conditions. Fig. 2-3 shows the five possible charge conditions in the collector. Note that these conditions are determined by the value of I_C and V_{BC} . This recognition is very important in developing the algorithm to derive the solution of the model equations.

Based on (2-2) through (2-5), the various modes of BJT operation were mapped on a V_{BC} - V_{BE} plane [Modified]. This plane of "model boundaries" then became the basis for the BJT model algorithm in the routine QMCT, into which values of V_{BE} and V_{BC} are passed. The model complexity of this mapping is less obvious problems with the algorithm.

The model boundaries can be depicted much more clearly on the I_C - V_{BC} plane of operation as implied by the above description. Fig. 2-4 shows the operational modes on the I_C - V_{BC} plane, with the boundaries labeled in accord with (2-1) through (2-5). In fact, the range of operation has been expanded beyond that as [Modified] based on insight afforded by this representation. This expansion is described in the next section.

The stored charge in the QBLs for cases (b) and (c) in Fig. 2-3 can be evaluated by using [Modified] and [Modified], respectively. Note that Q_{QBL} is the neutralized stored charge stored in the quasi-neutral *n*-p-collector region. Stored charge in the SCEs must also be evaluated, not only to define the transient model but also to physically link the base and collector analyses based on the total charge-neutrality imposed in the μm^2 region. The stored

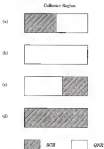


Fig. 3-6: Four possible charge conditions in the qpl collector



Fig. 2-4: Operational model boundaries in the $I_D - V_{DS}$ plane, where V_{DS1} is the value passed to the $\overline{MPT}$ model routine Q_{BCT} . See Fig. 2-7 for the points A to H.

charge in the RCH is derived analytically based on the model assumptions that the carrier velocity in the RCH is saturated at v_{sat} and that the carrier flow is predominantly by drift in the region. For cases (a), (c), or (d) in Fig. 3-5, the RCH charge is

$$Q_{\text{RCH}} = -qAW_{\text{RCH}} \left(\frac{I_{\text{C}}}{qAv_{\text{sat}}} + N_{\text{RCH}} \right) \quad \text{for } W_{\text{RCH}} \neq 1 \quad (2-6)$$

$$= -qAW_{\text{RCH}} \left(N_{\text{RCH}} + \frac{I_{\text{C}}}{qAv_{\text{sat}}} \right) \quad \text{for } W_{\text{RCH}} = 1 \quad (2-7)$$

When complete charge neutrality is imposed in the p - n base-collector structure, the charge stored in the buried layer, Q_{BL} , must be considered from a substantial electric field associated at the n -epi-collector buried layer junction. By Gauss's law,

$$Q_{\text{BL}} = eE_{\text{BL}}h \quad (2-8)$$

where E_{BL} is the magnitude of the electric field at the high-low junction in the collector region. In the case that $Q_{\text{RCH}} + Q_{\text{BL}} \neq 0$, Q_{BL} , which is the positive space charge in the base, should be included to satisfy the total charge-neutrality

$$Q_{\text{TC}} + Q_{\text{RCH}} + Q_{\text{BL}} = 0 \quad (2-9)$$

Next, the overall model algorithm will be reviewed qualitatively. The detailed description can be referred to in [Joshi04]. A deviation of the charge-based EFT model is that the model equations are solved numerically by marching the regional analysis through the moving boundary conditions. This is unlike the conventional Gauss-Poisson EFT model [Hettr01] which yields the solution analytically via explicit model equations.

Two primary regional analyses are carried out during each Newton-Raphson iteration of the current model analysis. The collector analysis first ascertains the position of the junction BCR corresponding to given I_C^0 and V_{BE} , where I_C^0 is a guessed or updated collector-current density assumed to flow in the opt collector region for specified V_{BE} and V_{BC} . If the junction BCR is found to exist, W_{BCR} , Q_{BCR} and Q_{BC} are calculated in turn. Q_{BC} is assumed to be depletion-charge of the base depends near the junction, hence the depletion-region edge x_D can be derived by integrating the base doping profile $N_A(x)$, which is assumed to be exponential in the model, from x_D to W_{BCR} into Eq. 3-11, and equating the integral to Q_{BC} . Then $N_A(x_D)$ is known. To complete the collector analysis for this case, the electron density at x_D is approximated as

$$n(x_D) = \frac{I_C}{qA_D} \quad (3-12)$$

The results x_D , $n(x_D)$, and $N_A(x_D)$ constitute moving boundary conditions for the base analysis.

If the junction BCR is found to be collapsed, as in quasi-saturation, the collector analysis follows a different procedure to obtain the boundary conditions. First, the QNL width is evaluated as $W_{BCN} - W_{BCR}$, where W_{BCR} is derived from the current-induced BCR analysis. Then Kallio model [Kall66] is applied to the QNL. The QPL separation $V_{BCN} = (E_{BP}/q) - E_{BP}(R1)/q$ at the collector base metallurgical junction can then be computed numerically from the expression

$$I_C^0 = \frac{E_1(V_{BCN}) - E_1(V_{BCR}) - \ln \left[\frac{1 + E_1(V_{BCN})}{1 + E_1(V_{BCR})} \right] + \frac{V_{BCN} - V_{BCR}}{V_T}}{R_{BP}/V_T} \quad (3-13)$$

where $K_T(T) = (1 + \exp(V_T/V_D))^{-1/2}$ with $V_T = (kT/n_f/M_{\text{eff}})^{1/2}$, $M_{\text{eff}} = W_{\text{gate}}/(2\phi_{\text{sc}}M_{\text{gate}})$, and $V_D = (kT)/q$. Here, V_{eff} is defined as the QFL separation at the transition edge between the QSE and the current-reduced SCE. In HSPICE 1.3, it is approximated as [Laffeta]

$$V_{\text{eff}} = V_D \ln \left[\frac{\frac{I_C}{qA\tau_2} \left(\frac{I_C}{qA\tau_2} - M_{\text{eff}} \right)}{n_C^2} \right] \quad (2-12)$$

under the assumption that the electron density is continuous at the transition edge and that quasi-neutrality prevails there. But the definition of V_{eff} is not physical and leads to many problems in the implementation of the model. An improved physical definition will be given later. The computed V_{eff} is used to derive the boundary conditions for the base analysis. Without the junction SCE, $n_C = W_{\text{DE}}$ since no depletion region exists. Thus $M_{\text{eff}}(n_C = W_{\text{DE}})$ is implicitly available, and $n(n_C = W_{\text{DE}})$ is derived under the quasi-neutral approximation from the expression

$$n(W_{\text{DE}}) (n(W_{\text{DE}}) + M_{\text{eff}}(W_{\text{DE}})) = n_C^2 \exp \left(\frac{qV_{\text{eff}}}{kT} \right) \quad (2-13)$$

These boundary values are passed to the base analysis. As implied in this discussion, the model tends to be less efficient in quasi-neutrality due to the additional iteration required to solve (2-13) for V_{eff} . A method to avoid this loop and significantly improve the computational efficiency will be described later in this chapter.

The base analysis, the assumed doping profile $N_A(x) = N_{\text{DE}} \exp(-qx/W_{\text{DE}})$, yields a collector current I_C^B with the boundary conditions defined from the collector analysis. It is calculated through a weighted sum of low and

high-frequency carrier transport solutions, in both the forward and reverse modes (Joshi)

$$I_0^{\pm} = \frac{I_{CLF} + C_F \frac{n_0}{W_{eff}} I_{CLF}}{1 + C_F \frac{n_0}{W_{eff}}} = \frac{I_{CLR} + C_R \frac{n_0}{W_{eff}} I_{CLR}}{1 + C_R \frac{n_0}{W_{eff}}} \quad (2-16)$$

where $I_{CLF} = (qAn_G)/(\tau^2/2) \cdot (1)$, $I_{CLR} = (qAn_G)/(\tau^2/2) \cdot (1)$, $I_{CLF} = (2An_G)/(\tau^2 \ln(f))$, and $I_{CLR} = (2An_G)/(\tau^2 \ln(f))$ with $n = (qphD_s \tau)/W_{eff}$, $f = \exp((\tau W_G)/W_{eff})$ with $W_G = x_G + x_D$, $n_G = n(x_G)$, $n_D = n(x_D)$, $N_{AG} = N_A(x_G)$, and $N_{AD} = N_A(x_D)$. Note that n_G , n_D , and $N_A(x_G)$ are functions of $V_{GG} = 0V$, and hence are calculated directly from an analysis of the smaller-bias junction IVCs (Joshi)

The basis for the device control is $|V_G^+ - V_G^-|$. An expanded second method, with separate bias-toning and bias-sweeping, is used until $|V_G^+ - V_G^-|$ converges over a specified minimum limit. Then the iteration stops, and the relevant expanded charges and the bias-current components are computed prior to exiting QICT and returning to the model analysis.

2.2 New Model Features for MMSPICE-1.2

2.2.1 Expanded Charge Expressions

Our initial approach to enhancing the model runtime will be to ensure the continuity of the regional charge expressions for each operational mode across all model boundaries. This will result in different (more physical) characterizations of some of the charges.

Regional charge expressions under various bias conditions are necessary to properly model the terminal dc currents as well as the transient behavior of the device. These expressions also make possible an accounting for non-quant-

static charge dynamics by charge partitioning (Pohl) in the charge-based NUT model, the regional charges are modeled directly with physical parameters, without using empirical transit times. We stress here that charge continuity across all the model boundaries in Fig. 3-4 is strictly required to ensure that the terminal currents are continuous, especially for steady-state analysis, as well to promote convergence. MMSPICE-1.2 is deficient in this regard.

In our derivation of the charge expressions, we will consistently assume that the electric field in the SCFs is not less than E_p . This implies that the space charge is regions where the field is less than E_p is neglected, unlike in MMSPICE-1.2 (see Fig. 3-4). In hard saturation, the entire eye-collector region is again neutral, and the field intensity is much less than E_p . Thus, no space charges are assumed. In this case, $Q_{\text{eye}}^{\text{eye}}$, which is the expected mean charge in the quasi-neutral eye-collector region associated with base push-out, must be characterized to explain the observed charge dynamics. Its expression can be derived following the works [Joshi] and [Kohli].

In addition to hard saturation, there exist four distinguishable modes of operation based on their different electric field patterns at the eye collector. To systematically represent the regional charges, the electric field patterns corresponding to each operational mode are taken into consideration. With reference to Fig. 3-4, the typical field patterns corresponding to operational points A to E are shown in Fig. 3-7, with some variations that will be referred to as the following throughout.

First, we derive the regional charge expressions for the mode C (see Figs. 3-4 and 3-7). In doing so, we will get a new expression for W_{eye} to replace (2-1) that ensures model consistency in the derivation of the charges. Under the approximation that the potential drop within the base and the biased layer can

be neglected, we obtain the new expression for $\bar{W}_{ACC}$ by integrating the electric field depicted in Fig. 3-1(a) across the entire *opt* collector layer and equating it to $\phi_C = V_{AC}$:

$$\bar{W}_{ACC} = -\frac{e\bar{R}_2}{4\pi\epsilon_0\epsilon_{eff}} + \sqrt{\left(\frac{e\bar{R}_2}{4\pi\epsilon_0\epsilon_{eff}}\right)^2 + \frac{2e\Delta\bar{R}_2}{E_C - E_G}(\phi_C - V_{AC} - I_C R_{ACC})}. \quad (2-48)$$

Note that the electric field is modeled to be discontinuous at the junction SCH edge $\bar{W}_{ACC}$, unlike in Fig. 3-1(b), in compliance with the aforementioned assumption. The significance of this modification is reflected by the resulting continuity of the charges at the metal boundary between (a) and (b) in Fig. 3-1.

By differentiating $\bar{W}_{ACC}$ with respect to I_C and equating the result to zero, we get the same expression for V_{ACC} as described previously. Also, the following results can be derived:

$$\bar{W}_{ACC} = 0 \quad \text{for } I_C > I_{ACC} = \frac{q_0 - V_{AC}}{R_{ACC}} \quad \text{and } V_{ACC} < V_{BO}$$

and

$$\bar{W}_{ACC} = \bar{W}_{AV} \quad \text{for } I_C < I_{ACC}/\epsilon_{eff} \quad \text{and } V_{AC} = V_{CMT},$$

where

$$I_{ACC/eff} = \left(I_C + \frac{V_{AC} - V_{CMT}}{R_{ACC}} \right), \quad (2-49)$$

which replaces $I_{ACC/eff}$ in (2-10). With (2-45), Q_{B21} corresponding to Fig. 3-1(a) is modeled by

$$Q_{B21} = q\delta_C \left(N_{AV} + \frac{I_C}{q\delta_C V_{AC}} \right) W_{ACC} \quad \text{for } \bar{W}_{ACC} > 0 \quad (2-50)$$

The buried-layer charge Q_{BL} , expressed by (3-8), is neglected in this mode since R_{BL} is always less than R_T . The total charge negatively three decades Q_{GRT} is to be balanced with Q_{GCR} , which is the total junction space charge in the base

$$Q_{GCR} = -(Q_{GRT} + Q_{BL}) = -Q_{GRT} \quad (3-10)$$

For quasi-neutral regions containing excess carriers solely in this case, then $Q_{GRT} = 0$.

Second, the regional charges for mode (1) in Figs. 3-4 and 3-5 will be determined. This mode is called *non-charged quasi-injection*. As can be seen in Fig. 3-5(a), the junction SCR is collapsed. By applying Gauss's law, under the model assumptions for this mode, Q_{GRT} and Q_{BL} can be derived as

$$Q_{GRT} = -qA \left(\frac{L_T}{2\pi W_{GRT}} - W_{GRT} \right) W_{GCR} \quad (3-11)$$

with W_{GCR} given in (3-6), and

$$Q_{BL} = -Q_{GRT} \quad (3-12)$$

in this case Q_{GCR} is zero since there exists no junction SCR. For the basic quasi-neutral n-p region, Q_{GCR} should be expressed properly. Applying Hall's model [Hall(5)] but extending it further to account for the reduced QPN width, Q_{GCR} can be expressed as a function of $W_{GCR} = W_{GRT} - W_{GCR}$ and V_{GRT} , which is defined at the transition edge, as well as V_{GCR} . However, a newly defined V_{GCR} , which will be discussed in the next section, is used instead of (3-13).

Third, the charge quasi-injection mode corresponding to the electric field patterns shown in Fig. 3-5(b) and mode (b) in Fig. 3-5 will be considered. The emitter-collector layer is divided into two electrically different regions in this

case, one is the conductivity-modulated quasi-neutral region adjacent to the junction and the other is the ohmic neutral region. Until I_C reaches I_D , the ohmic region represented by Ψ_{ohmic} in Fig. 3-7(b) supports the entire potential drop $\phi_p = V_{BC}$. No space charge is induced in this mode of operation. Q_{BL} can be neglected since $E_{BL} \ll E_D$. Obviously, Q_{BL} cannot be formed since the electric field is less than E_D everywhere in the npl collector, hence $Q_{BL} = 0$. To account for the conductivity modulation, Q_{BL} must be expressed with the parameters that describe the modulated width and the degree of asymmetry of the current. It is expressed in terms of $\Psi_{BET} = \Psi_{\text{BETON}} - V_{BET}$ and V_{BCO} using the modified Hall analysis, as in the non ohmic quasi-saturation mode.

Fourth, mode III of Fig. 3-4 (corresponding to Fig. 3-7(d)) is discussed. As the electric field pattern shows, the entire npl collector is space-charged, but the mode can be separated into two different regions according to the region of the space charge. If $I_C > I_D$, which corresponds to G and H in Fig. 3-7(d), the current-induced SCR is formed, if $I_C < I_D$, which corresponds to D and E in the same figure, the depleted junction SCR, which is partially compensated by the injected electrons, is formed. At the onset current I_D , the electric field becomes flat as denoted by F in the figure, and no space charge exists. In all cases, the field is not less than E_D anywhere in the npl collector. The space charge Q_{BL} in this mode can be expressed as

$$Q_{BL} = qA \left(M_{BET} - \frac{I_C}{qAE_D} \right) = \frac{W_{BET}}{E_D} (I_D - I_C) \quad (2-21)$$

And, since E_{BL} can be expressed explicitly as

$$E_{BL} = \frac{\phi_p - V_{BC}}{E_{BET}} = \frac{qW_{BET}}{E_D} \left(M_{BET} - \frac{I_C}{qAE_D} \right) \quad (2-22)$$

(2-22) yields

$$\begin{aligned}
 Q_{\text{BL}} &= -kA(E_{\text{BL}} - E_D) \\
 &= \frac{kA}{W_{\text{JVT}}} (V_D - V_{\text{BL}}) = \frac{kA W_{\text{JVT}}}{L} \left(N_{\text{BLT}} - \frac{I_C}{qA v_{\text{th}}} \right) = -kA E_D
 \end{aligned}
 \tag{2-30}$$

where the last term complies with our assumption that the space charge is neglected if the field is less than E_D . From Q_{JVT} and Q_{BL} , Q_{JCT} is obtained by imposing the total charge neutrality:

$$Q_{\text{JCT}} = -(Q_{\text{JVT}} + Q_{\text{BL}}) \tag{2-31}$$

Mode (d) is actually a mixture of (a) and (c) depending on the current level, *i.e.*, it cannot be said to be strictly a forward active or a quasi-saturation mode. The important aspect of this mode is the presence of the SCB at the junction, which violates the linkage of the base and collector depletion.

In mode (d), a noteworthy operational condition must be pointed out. From (2-10) and (2-14), W_{JCT} can be W_{JVT} even if $I_C \neq 0$. This occurs when $V_{\text{BC}} < V_{\text{BCJVT}}$, where V_{BCJVT} is the base-to-collector voltage defined as $V_{\text{JCT}} = I_C R_{\text{JCT}}$, at which (2-10) becomes zero, as depicted in Fig. 2-4. This finding is significant, and actually resulted in an additional major branching in the new device model, as mentioned in Section 2.1.4. This mode was ignored in MMSICE-1.2.

The regional charges for each mode derived so far are summarized in Table 2-1. With reference to Fig. 2-4 and this table, it can be verified that the charges are indeed continuous at all the modal boundaries. These results come entirely from the consistent derivation of the charges under the model assumptions, and they preclude the need for smoothing functions used in MMSICE-1.2.

Table 2-4. Regional Charge Expressions

region charge distribution mode	Q_{reg}	Q_{res}	Q_{gen}
(a)	$\frac{W_{\text{ACC}}}{v_g} (I_C - I_C^0)$	0	0
(b)	0	0	$\frac{(1 - W_{\text{ACC}}) W_{\text{gen}} \left[\exp \left(\frac{V_{\text{ACC}}}{V_T} \right) - \exp \left(\frac{V_{\text{gen}}}{V_T} \right) \right]}{(1 - W_{\text{ACC}}) W_{\text{gen}} + W_T}$
(c)	$\frac{W_{\text{ACC}}}{v_g} (I_C - I_C^0)$	$-Q_{\text{gen}}$	same expression as above
(d)	$\frac{W_{\text{ACC}}}{v_g} (I_T - I_C^0)$	$\frac{dI}{dV_{\text{gen}}} (I_C - V_{\text{gen}})$ $-\frac{W_{\text{gen}}}{2v_g} (I_C - I_C^0)$ $-dI/dV_T$	0

Note: $Q_{\text{ACC}} = -Q_{\text{gen}} + Q_{\text{ACC}}$

3.1.3 Quantum Level Separation, V_{AC}

For quasi-equilibrium operational modes, V_{AC} , previously given by (2.12), is needed in the collector analysis. In MINIPACK 1.3, it is modeled only for the one-dimensional quasi-equilibrium mode (a) under the assumption that charge neutrality is valid and the electron carrier density is continuous at the beam exit point $x = W_{gap}$ (see Fig. 2.7). This assumption, however, is not physical, and it causes problems in the quasi-equilibrium mode (a) where the argument of the natural logarithm in (2.12) is negative. This difficulty necessitated the use of an empirical expression for V_{AC} in the mode (a) analysis, which unavoidably causes formidable discontinuity problems across the model boundaries that occurred mode (b) in Fig. 2.4.

We can physically define V_{AC} with the consideration that for quasi-equilibrium conditions the junction barrier has vanished, and the total potential drop across the unmodulated drift gap region (or quasi-equilibrium across the current-induced SCE across-drift quasi-equilibrium) is $\phi_C = V_{AC}$. With the aid of the band diagram shown in Fig. 2.4, V_{AC} can be shown to equal ϕ_C . The QFL separation at the high-low junction between the gap collector and the biased layer is

$$E_{CE}(W_{AC}) - E_{CP}(0) = qV_{AC} \quad (2-25)$$

and, in accord with the above statement,

$$E_C(W_{gap}) - E_C(W_{AC}) = q(\phi_C - V_{AC}), \quad (2-26)$$

if we neglect the potential drop across the transition region and across the biased layer. The electron carrier density within the unmodulated drift region or the current-induced SCE is constant, equal to N_{AC} or $i_C/(q\mu_n E_C)$.

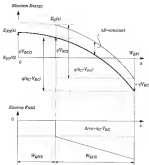


Fig. 1-6 Energy-band diagram and corresponding electric field profile included in the open collector region. The transition region between the QNR and the NCR is assumed to be negligible.

respectively. Thus, $E_y(x) = E_{DDE}(x)$ is constant within the mentioned regions, i.e., $E_C(x)$ is parallel to $E_{DDE}(x)$. Therefore, with (2-24) and (2-25) we directly infer from Fig. 2-4 that

$$\begin{aligned} qV_{DDE} &= E_{DDE}(W_{DDE}) - E_{DDE}(0) \\ &= [E_{DDE}(W_{DDE}) - E_{DDE}(W_{DDE})] + E_{DDE}(W_{DDE}) - E_{DDE}(0) \\ &= [E_C(W_{DDE}) - E_C(W_{DDE})] + E_{DDE}(W_{DDE}) - E_{DDE}(0) \\ &= q\phi_D. \end{aligned} \quad (2-27)$$

This definition is physical and indeed resolves all the problems that occur in MINIMOS-1.2 at the metal boundaries surrounding node (a) in Fig. 2-4. Further, this definition leads to the proper characterization of V_{DDE} , which must be greater than the built-in potential ϕ_D when the positive SCE has vanished in quasi-equilibrium.

3.3.3 Electron Carrier Density $n(x_D) = n_C$

The expression for the collector current i_c (3-14) from the base analysis depends on $n(x_D)$, the electron density at the collector-side edge x_D of the quasi-control base region. For quasi-equilibrium conditions, this density is defined by $V_{DDE}(x_D) = W_{DDE}$ by (2-12). However, the density is most difficult to define when the positive SCE exists due to the complexity of the underlying physics at the edge of the region. MINIMOS-1.2 assumes $n(x_D)$ for this case to be equal to the velocity-saturated carrier density in the positive SCE as given in (2-10). This approximation, however, is not acceptable because it leads to discontinuity of $n(x_D)$ and thus I_C at the metal boundaries.

Physically, $\psi(x_Q)$ should be defined in terms of the QFL separation at the edge of the SCE.

$$\begin{aligned} \psi(x_Q) (\psi(x_Q) + E_A(x_Q)) &= \kappa_Q^2(x_Q) \left(\frac{E_{xx}(x_Q) - E_{xx}(x_Q)}{q^2} \right), \quad (3.48) \\ &= \kappa_Q^2(x_Q) \left(\frac{E_{xx}(x_Q) - E_{xx}(x)}{q^2} \right), \end{aligned}$$

where it is assumed that the hole quasi-Fermi level is constant through the SCE. From the collector analysis, κ_Q is obtained explicitly from Q_{AC} . But $E_{xx}(x_Q) - E_{xx}(x)$, which depends on the unipolar transport in the quasi-neutral base, is not available in a closed form for modes (a) and (b) in Figs. 3-3 and 3-5. $E_{xx}(x_Q)$ is generally a function of V_E and V_{BC} . To overcome this difficulty, and to model $\psi(x_Q)$ properly, an expression for $E_{xx}(x_Q)$ which can reasonably account for the complex physics is needed.

To characterize $E_{xx}(x_Q)$, three physically reasonable assumptions are made. First, $E_{xx}(x_Q)$ is assumed to be $E_{xx}(x) - qV_{BC}$ (see Fig. 3-3) for the case of the zero collector current. Thus, the QFL separation at the depletion edge is $-qV_{BC}$ in this case. This approximation is not accurate in BJT modeling. Second, $E_{xx}(x_Q)$ is assumed equal to $E_{xx}(x) + qV_E$ at the collector current I_{CCEsat} and I_{CC} (the threshold currents at which the base push-out effects occur) to assure that the QFL separation is continuous at all model boundaries. This approximation is consistent with the previous discussion on V_{BC} . Finally, the quasi-Fermi level separation at the depletion edge is assumed to vary logarithmically with the collector current I_C . This argument can be partially supported by the integral charge-control relationship [Mittl] its validity can be also supported by the fact that the first derivatives of the terminal variables (e.g., the collector current I_C) with respect to the indepen-

fixed terminal voltages (e.g., the collector-emitter base voltage V_{CE}) are continuous and reasonably well defined at the model boundaries.

The above three approximations lead to a semi-physical expression needed to define $I_{BES}(V_{CE})$

$$\exp\left(\frac{qV_{BES}}{kT}\right) = \exp\left(\frac{qV_{BE}}{kT}\right) + \left[\exp\left(\frac{qV_{BE}}{kT}\right) - \exp\left(\frac{qV_{CE}}{kT}\right)\right] \frac{I_C}{I_{CEFF}} \quad (2.28)$$

where $V_{BES} = [I_{BES}(V_{CE}) - I_{BES}(0)]/q$ and denotes the quasi-Fermi potential separation at the base-side edge of the collector junction SCR, and I_{CEFF} denotes the critical boundary current $I_{BESCRIT}$ or I_{BES} depending on V_{BE} relative to V_{CEFF} (see Fig. 2-8). With (2.28), the previous discontinuity of $n(x_E)$ at the model boundaries is removed completely, and the convergence problems originating from the discontinuity are substantially improved.

2.2.6 Modified Newton Control Scheme

The charge-based BJT model implemented in QBCV of MINISPACE consists of the regional analysis, which iteratively constructs a numerical Newton to solve the model equations. To make clear our discussion on a new algorithm, a brief qualitative review of the old one used in MINISPACE 1.3 will be given.

As discussed in Section 2.2, the collector analysis yields a set of boundary conditions, for given I_{B1}^0 , to be used in the base analysis, which in turn predicts I_{C1}^0 with the provided boundary conditions. The difference between I_{C1}^0 and I_{B1}^0 controls the next iteration, and the new I_{B1}^0 is defined through a monotonic-like method (bisection). The iteration continues until $|I_{C1}^0 - I_{B1}^0|$ converges within a specified tolerance. When the iteration converges to the solution (I_{C1}^0) ,

the regional charges are finally computed and partitioned among physically between the emitter and collector terminals to define Q_{BE} and Q_{BC} . Then, the base current components that constitute I_B are calculated.

The collector analysis is subdivided into several branches to cope with the various modes of operation. For quasi-saturation, viz. (3) and (4) of Figs. 2-5 and 2-7, an inner-loop numerical iteration is needed to evaluate V_{BCO} by deriving the reverse solution of (2-11). This observation helps us understand why the MINICIDE-1.1 model routine tends to be computationally less efficient in the quasi-saturation regime. The inner-loop for V_{BCO} can be avoided, however, by refinement of the outer-loop iterative control. Such refinement is now discussed.

The new algorithm switches the start of the regional analysis from the collector to the base for the case of quasi-saturation to avoid the unnecessary inner-loop iteration. To understand the rationale for the switching of the analysis, it will be helpful to consider the behavior of I_C^R as a function of I_C^F (or I_C^L) as a function of I_C^R , realizing that the solution we seek is

$$F_1 = I_C^F - I_C^R = 0. \quad (2-20)$$

Since the solution must exist, the general tendency of I_C^R with increasing I_C^F can be depicted as in Fig. 2-9(a) for fixed V_{BE} and V_{BC} . Or if I_C^L is depicted with increasing I_C^R under the same biases, it tends to be as in Fig. 2-9(b). This observation enables us to predict where the solution is relative to the current levels (given by substituting I_C^F & I_C^L as given for the zeros). For example, if I_C^L , which can be evaluated under the boundary conditions that are calculated with the guessed I_C^F in the collector analysis, is less than I_C^R ($F_1 < 0$), then the solution of (2-20) can be seen in Fig. 2-9(a) to be below I_C^F . The argument is reversed if I_C^L is greater than I_C^R ($F_1 > 0$), in which case there below the given I_C^F .

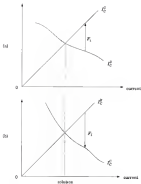


Fig. 3-5 The Gauss-Seidel iterative scheme: (a) K_1^0 is guessed first, and then K_2^0 is substituted; (b) vice versa.

This consideration shows that a specific value of I_C^0 can be chosen as the initial guess which will enable the determination of whether or not the junction SCR exists at the solution current. If it does not, then the base analysis can be done prior to the collector analysis avoid the additional inner-loop iterations mentioned above. In Fig. 2-4, the base models for the initial guess current is the boundary current that divides modes (c) and (d) or (a) and (b), i.e., I_{BCEQ2} or I_{BCE} depending on V_{BC} . If the solution current I_C^0 then is less than I_{BCEQ2} or I_{BCE} , it can be argued that there exists a junction SCR, if it is greater, then the junction SCR does not exist, and quasi-saturation prevails.

Based on this insight, a new model algorithm was developed as illustrated in Fig. 2-10. During the first iteration, I_C^0 is evaluated using the aforementioned boundary current for I_C^0 as the initial guess. A positive value of F_1 then implies that the solution lies in a forward-active mode in which a finite junction SCR exists, then, MFLAG is set to 1. A negative value of F_1 indicates that the solution falls into a quasi-saturation mode, then, MFLAG is set to -1. In the subsequent iterations, the base analysis starts first, and then the collector analysis follows if MFLAG=1, and vice versa (as before) if MFLAG=-1. In each iteration, $F_1 = I_C^1 - I_C^0$ is checked, and according to its value, I_C^0 is updated for the next iteration via the secant-bisection method if MFLAG=1, or, I_C^0 is updated if MFLAG=-1. The iteration stops when $|F_1| < \text{ABSTOL}$, which is one of the optional parameters that control the convergence in SPICE. Thus, the key points in the new iterative control are the change of the initial guess for I_C^0 to determine during the first iteration if the solution corresponds to a mode of operation in which the junction SCR exists or not, and the switch of the regional analysis order in the case of quasi-saturation (can evaluate V_{BC} directly from the base analysis, thereby eliminating the inner-loop iterations in the collector analysis).

To clarify how the beam analysis is done first, to remove the inner loop, consider $\mathcal{L}_C^B$ given in (2-10). With no junction-SCB sub-space variations, which means $\alpha_C = V_{BIB}$, (2-10) is a quadratic equation with respect to $\alpha(\alpha_C)$. Then, $\alpha(\alpha_C = V_{BIB})$ can be derived explicitly if $\mathcal{L}_C^B$ is given, since all the other parameters are determined. Then, V_{BIB} is calculated from (2-10) straightforwardly, and is passed to the collector analysis to evaluate $\mathcal{L}_C^B$. The above procedure to obtain V_{BIB} is done only when the junction SCB does not exist. Otherwise the original sequence of analyses is done.

In calculating $\mathcal{L}_C^B$ with V_{BIB} , it is required to evaluate R_{BIB} in (2-11), which is also a function of the collector current to be solved. In MMSPICE-1.1, it is evaluated with $\mathcal{L}_C^B$, unlike in MMSPICE-1.3, where it is estimated with $\mathcal{L}_C^B$.

The new model algorithm in Fig. 2-18 furthermore modifies refinement and expansion of the collector analysis as implied by the new modeling described in Sections 2.3.1-2.3.3. In accord with the revised operational mode characterization in Fig. 2-6, the collector analysis now comprises four branches ($V_{CB} < -V_{CB}$, $-V_{CB} < V_{CB} < -V_{CBIB}$, $-V_{CBIB} < V_{CB} < -V_{BIBBIB}$, and $V_{CB} > -V_{BIBBIB}$), each containing sub-branches defined by $\mathcal{L}_C^B$ relative to model-boundary currents. Note that now, since $\alpha(\alpha_C)$ is defined more physically, the reverse mode where $\mathcal{L}_C^B < 0$ is properly accounted for. MMSPICE-1.2 does not do this, nor does it correctly account for the case $V_{CB} = -V_{BIBBIB}$ mode, where the op-collector is completely space-charged independent of $\mathcal{L}_C^B$, as does MMSPICE-1.3. Note that for each and every sub-branch analysis, the algorithm is dictated by the particular charge condition in the collector as indicated in Fig. 2-6.

The modification of the device control, with the aforementioned improvements in the model itself, implemented in predominantly new codes

code for QMCT in MONSPICE-1.3. With the modification, the computational efficiency of the model routine is improved greatly in quasi-stationary modes, and is now comparable to that in linear-stationary modes. Some simulation results are presented in the next section to demonstrate the improvement.

3.4 Simulation Results

The overall performance of MONSPICE-1.3 is superior to MONSPICE-1.2 in every respect. The convergence problems are substantially reduced, and the consistency of the derivatives of the terminal characteristics gives MONSPICE significant potential as an analog circuit simulation tool. To demonstrate the improved performance, some results are benchmarked.

Fig. 3-11 shows the dc output characteristic of a MOSFET-III (Silvaco) BJT as simulated with MONSPICE-1.3 and contrasted with the MONSPICE-1.2 simulation. MONSPICE-1.2 gives a flat I_C (satiate r_C) within some regions of bias; this non-physical result does not occur with MONSPICE-1.3. In Fig. 3-12, the Early voltage, $V_A = I_C / (dI_C / dV_{CE}) = V_{CE}$, derived from the predicted characteristic in Fig. 3-11 is plotted as a function of V_{CE} . MONSPICE-1.2 yields a spurious discontinuity at V_A ($r_C = \infty \Rightarrow V_A = \infty$), which limits its utility as an analog circuit simulator. MONSPICE-1.3 shows a dramatic improvement, predicting a monotonic increase in V_A with V_{CE} , which is physical (and which is unaccounted for in the Gummel-Poon BJT model [Gat70]).

To demonstrate the improved computational efficiency in quasi-stationary or stationary of MONSPICE-1.3, the total job times for the dc simulations of Fig. 3-11 are given in Table 29. For each value of I_C , simulations were done using 0.5-V steps in V_{CE} . As reflected by the increasing run times of MONSPICE-1.2 for increasing I_C , most of the simulated operating conditions fall into quasi-stationary modes. However, the results of MONSPICE-1.3 show

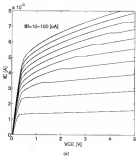


Fig. 3-12 Predicted dc output characteristics of a MOSAIG-3 (Silix) JFET with junction area of $20 \mu\text{m}^2$ (a) simulated with MINIMOS-3.2, (b) simulated with MINIMOS-3.2

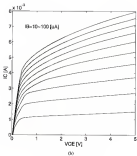


Fig. 1.13—Continued.

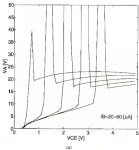


Fig. 3-12 Early voltage plots corresponding to the simulated results of (a) MMSF100-10 and (b) MMSF100-1 shown in Fig. 3-11.

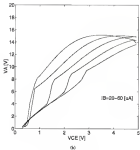


Fig. 9-75-Continued

Table B3: DC Analysis Total Job Time Comparison

Item number I_p	Lot Lot	MMOPCE 1.1 (sec)	MMOPCE 1.2 (sec)	Remarks
1		1.58	1.58	FA
2		(NG)	1.58	
3		2.37	1.58	
4		2.58	1.58	
5		2.45	1.58	
6		(NG)	1.58	
7		2.52	1.58	
8		1.88	1.58	
9		1.65	1.58	
10		(NG)	1.58	F
15		2.13	1.58	QS
20		3.83	1.43	
40		3.83	1.53	
50		(NG)	1.77	
60		4.47	1.35	
70		4.63	1.45	
80		4.63	1.73	
90		5.35	1.43	
100		5.85	1.40	F

Note: FA = Forward Action, QS = Queue Substitution, NG = Not Converged

only a minimal increase in run time with I_{D0} , demonstrating the efficiency of the new QMOT device model, and indeed the enhanced utility of MMSPICE.

The convergence problems are also substantially reduced as indicated in Table 3.3. As long as most of the transistors simulated in a circuit are not in cut-off modes, it is found that the convergence problems are virtually eliminated in MMSPICE-1.3. Any non-convergence may be easily removed by simply using reasonable initial (V_{DD} and V_{SS}) values for the devices, but this was not the case with the old version. The remaining convergence problem, which tends to occur at very low-current (cut-off) conditions in BJTs, seems to be a SPICE3 problem related to large values of $\Delta BDTOL$. Increasing $\Delta BDTOL$ above the default value (10^{-10} A) can improve the convergence, but at some sacrifice in accuracy at low-current conditions.

As a benchmark, a four-stage KTL inverter circuit was simulated in the same (transient) domain on a SUN-4 SPARC machine. The simulated time interval was from zero to 200 nano-seconds with 1 nano-second time step. The input pulse with voltage swing from zero to 4 was assumed to have rising and falling times both of 1 nano-second, and period of 100 nano-seconds. The supply voltage was 5. The MMSPICE input file is shown in Fig. 3-16. This circuit operates mostly in-quiescent evaluation and cut-off modes, and thus also can demonstrate the improved computational efficiency of the new device model algorithm. The total job-times were 35 sec and 18 sec with MMSPICE-1.3 and MMSPICE-1.3, respectively. This dramatic improvement reveals well the superiority of MMSPICE-1.3, and indeed indicates that the substantive benefits offered by its physical modeling mechanism had by relatively little cost in CPU time over that required by SPICE3 and its empirical BJT model (Shafiq). The general conclusion is confirming the usefulness of the more mechanistic models used in *Profilis*, which include a Colpitts oscillator, a MOSCL-01 KCL,

Abstract **INTRODUCTION:** The purpose of this study was to determine the effect of a 12-week, supervised, low-impact, aquatic exercise program on the physical and psychological health of older adults with arthritis. **METHODS:** A total of 100 subjects with arthritis were recruited from a local senior center and a local arthritis support group. The subjects were randomly assigned to either the exercise group or the control group. The exercise group performed a 12-week, supervised, low-impact, aquatic exercise program. The control group did not participate in any exercise program. **RESULTS:** The exercise group showed significant improvements in physical and psychological health compared to the control group. **CONCLUSIONS:** A 12-week, supervised, low-impact, aquatic exercise program can improve the physical and psychological health of older adults with arthritis. **KEYWORDS:** arthritis, exercise, low-impact, aquatic, older adults.

```

VIN 1 0 DC 1 0VRL0 4
+ 100u 4 500u 5 600u 5 700u 4 800u 4
+ 1100u 4 1200u 5 1300u 5 1700u 4 2000u 4)
VDC 10 0 DC 5 0

BC1 10 0 0 0u
BC2 1 0 0 0u
Z1 0 0 0 0 NPNMOD AB=10p AC=10p WB=10u TC=0 0 4 0
BC3 10 0 0 0u
Z2 0 0 0 0u
Z3 0 4 0 0 NPNMOD AB=10p AC=10p WB=10u TC=0 0 0 0
BC4 10 1 0 0u
Z4 0 0 0 0u
Z5 1 0 0 0 NPNMOD AB=10p AC=10p WB=10u TC=0 0 4 0
BC6 10 0 0 0u
Z6 0 1 0 0u
Z7 0 0 0 0 NPNMOD AB=10p AC=10p WB=10u TC=0 0 0 0
CL 0 0 0 1p

.MODEL NPNMOD GQPNP
+ JEHP= 1000 HEP=2 5000 10 WPT=0 1000 0 WTM=0 1000 0
+ KTA=0 1 MA=0 4045 10 TC=0 1000 0 TB=0 5000 50
+ TE=0 1000 10 JED=1 0 0 0 0 JEDC=1 1000 00 MOC=0 0
+ WSED=0 0 0 0 0 0 PMA=0 0 0 0 0 0 0 0 0 0 0 0
+ DVB=0 0 0 0 0 0 0 0 0 0 0 0 0 0 0 0 0 0
+ CBE=0 0 0 0 0 0 0 0 0 0 0 0 0 0 0 0 0 0
+ ICE=0 0 0 0 0 0 0 0 0 0 0 0 0 0 0 0 0 0

.TRAN 10 5000
PRINT TRAN VIB
PLOT Q10 0 0
END

```

gate, an unstable multibit device, and a voltage-controlled oscillator, as well as the RTL circuit. The comparisons of MMSPICE-1.3 versus MMSPICE-1.2 simulations show in all cases the superiority of the new version regarding convergence, strategy and computational efficiency.

2.5 Summary

A unified and physical approach for modeling the advanced EBT for all injection levels has led to MMSPICE. As an application specific mass-market device/device simulator, MMSPICE provides bipolar digital circuit designers with the capability of efficiently predicting device and circuit behavior down to the device structure, without requiring any empirical parameter-extraction procedure.

However, the first released version, MMSPICE-1.1, has several serious drawbacks in its formulation, which caused problems such as convergence difficulties and unrealistic predictions of small-signal device characteristics, like the unbounded Early voltage. Furthermore, it tends to be computationally less efficient when the transistor enters into quasi-saturation or subthreshold. MMSPICE-1.3 was evolved to remove these shortcomings, and possibly to extend the application range even to analog circuit simulation.

For the development of MMSPICE-1.3, substantial modifications were made in four areas. First, the regional charge expressions were modified especially to remove discontinuities across the operational modal boundaries. New expressions for W_{DEC} and $I_{DECINGR}$, based on the assumption that charge should be ignored within space-charge regions unless the electric field intensity is less than the critical field E_c , underlay the improved modeling. Second, the QFL separation V_{DEC} , which is defined at the transition region that is formed when the EBT goes into charge or non-charge quasi-saturation, was

newly defined based on the actual device physics. The expressions for Q_{gate} were then modified since it includes a V_{gate} term. Third, the characterization of the electron carrier density n (n_C) at the base-side edge of the collector junction (CJ) was substantially changed with the introduction of a semi-physical equation for the QFL separation at the junction, arranged to be a function of V_C and V_{BC} . Fourth, the steady-state scheme in QTCF was modified to close the lower circuit loop for the calculation of V_{BCQ} in quasi-stationary, and hence to make MMSFICE more computationally efficient. As a result of all these modifications, the entire model algorithm in QTCF was re-defined, and new source code for MMSFICE-1.3 was written.

To benchmark the performance improvement, several bipolar circuits were simulated, and results were compared with corresponding ones for MMSFICE-0.2. The overall performance of MMSFICE-1.3 was shown to be substantially superior, especially regarding convergence and computational efficiency.

MMSFICE-1.3 still has weaknesses in the hard saturation model for application to digital circuit simulations, and in the universal active mode model for application to analog circuit simulations. Also, there remain some convergence problems in simulating BJT's that are cut-off, or nearly so. This problem with low current convergence, however, seems to be SPICE3-related. Overall, the present status of MMSFICE-1.3 as an application specific, semi-empirical mixed-mode TCAD tool is reasonably good.

CHAPTER 3 SiGe-BASE HBT MODELLING FOR CIRCUIT SIMULATION

3.1 Introduction

Recently, the SiGe-base heterojunction bipolar transistor (HBT) has been actively investigated as a viable technology since it allows extreme vertical scaling without excessive base resistance for high-speed f_T and f_{MAX} performance [Craigh, Ghadi, Gohari, Haeber, Haeber, Haeber, Haeber, Haeber]. Extremely high current gain achieved in this technology [Schiff] provides effective trade-off with the emitter-base junction capacitance to enhance device speed. The energy bandgap across the quasi-neutral base region optimally graded with the incorporation of Ge can aid the minority carrier transport and hence further enhance device speed [Petroff].

Meaningful assessment of the (npn) SiGe-base HBT technology, and IC design based on it, will require a physical device model for circuit simulation that properly accounts for the nonuniform energy gap in the base, as well as the nonuniform doping density, both of which influence the electron transit time and hence the device speed. Empirical models based on some average energy gap [Kaufh], or effective Ge/mol fraction, in the base may represent the electron current adequately, but they cannot simulate the stored charge dynamics (which underlies speed performance) accurately.

We present in this chapter a physical but compact model for the SiGe-base HBT which accounts for the mentioned nonuniformities. The base is only as narrow as exponential doping profile in the quasi-neutral base (QNB) and

a linear bandgap, both of which are representative of an actual device. In fact, since energy-gap narrowing in highly doped silicon tends to vary logarithmically with respect to the doping density, the analysis enables verification of the HBT model and that for the scaled Si-BJT with significant energy-gap narrowing in the base. Assuming small hole current in the npn device, we describe the effective field acting on the electrons. Then integrating the electron current equation across the quasi-neutral base region for both cases of low and high injection, we express the collector current in closed-form terms of the boundary conditions at the moving emitter and collector sides of the QNE. Unlike the Generalized BJT model for which empirical transit times are needed to relate stored charge to current, the analysis also enables integration of electron density over the QNE to characterize the charge directly.

The base model is physically linked to the analysis of the carrier transport in the np-collector region described in Chapter 2, which physically accounts for quasi-saturation (e.g., base widening). And we extend the collector region analysis to model energy barrier formation in the conduction band at high currents with bandgap discontinuity at the collector junction. The parasitic energy barrier formation in the conduction band, which occurs at high currents at the collector heterojunction (CvHJ, YcHJ), causes many adverse effects such as excess carrier accumulation in the base and the increase of the base transit time to degrade device speed. The base current also increases abnormally due to the rapid increase of the accumulated charge in the base. These effects have been accounted for in device simulators, but not yet in any circuit simulator. We give a physical model for the barrier formation at each base point. The composite carrier transport analysis, which involves iterative solution, forms the basis for the complete physical (and predictive) charge-based DBCT model, which is implemented directly in SPICE2G-4.

This chapter is organized as follows. In section 3.2, we develop the comprehensive SiGe -base HBT model equations, which account for the bandgap grading in the base region in terms of moving boundary conditions. In section 3.3, an analytic model for the energy barrier formation at the collector heterojunction is developed. And in section 3.4, we derive the moving boundary conditions, which are affected by the energy barrier formation, to link the collector analysis to the base region analysis. In section 3.5, the complete charge-based, large-signal HBT model is summarized by discussing its network representation. This discussion includes a review of the base current components and an update of the recombination charges, all of which derive from the mean iterative solution of the base and collector analysis. New insight concerning proper partitioning of various charges between the emitter and collector terminals is also given. And in section 3.6, we conclude this chapter with a summary.

3.2 Model Development

3.2.1 Model Assumptions

The fundamental approximation in our modeling is that the conduction band has an discontinuity at the Si/SiGe junction. The electron affinity of SiGe alloy is close to that of silicon for the practical range of Ge mole fraction (roughly below 20%) [Fertis, Kozlita] although the practical maximum mole fraction depends on the maximum allowable thickness of the strained $\text{Si}_{1-x}\text{Ge}_x$ layer. Thus the energy band split that occurs in III-V or II-IV compound heterojunctions is not considered, but the valence band discontinuity is accounted for. This approximation allows the use of drift-diffusion model for the carrier transport across the heterojunction.

The alignment of the electron and valence junctions is an important issue, especially, misalignment can result because of different diffusivities of Ge and commonly used base dopants, e.g., boron [Pfeiffer, 2001a]. The junction misalignment can induce a permanent energy barrier at the recombination band, which results in minority carrier accumulation in the base region area at low currents, and thus deteriorates device speed. We consider here only the devices with well matched junctions, which would be the case of optimal and viable device design.

A one-dimensional (1D) *SiGe* base HBT structure is assumed in our model. The representative base doping profile of advanced HBT devices is represented with an exponential function, as in MINYCE-1.3,

$$N_A(x) = N_{AB} \exp\left(-\frac{x}{W_{AB}}\right) \quad (3-1)$$

where W_{AB} and n_i are the metallurgical base region width and the field factor, respectively.

The graded bandgap across the quasi-neutral base (QNB) region can generate an aiding built-in field to the minority carrier transport [Kroth] and hence reduce the base transit time significantly. We consider only a linearly graded bandgap because of two practical reasons. First, the bandgap reduction is proportional to Ge mole fraction, which is normally graded linearly across the QNB. Second, since the bandgap narrowing due to heavy doping is a logarithmic function of doping density, which is well represented with the exponential doping profile in (3-1), it becomes a linear function of position.

The bandgap reduction due to the small Ge mole fraction x , which appears mostly in the valence band, is given by [Pfeiffer]

$$\Delta E_V(x) = 0.54x \text{ (eV)} \quad (3-2)$$

and we assume $K(x)$ is linear. The bandgap narrowing due to the heavy doping can be accounted for by Shockley's model [33,34]

$$\Delta E_g(x) = 3kT \left[\ln \left(\frac{N_A(x)}{10^{17}} \right) + 1 \right] + 10kT \left(\frac{N_A(x)}{10^{17}} \right) + 0.7kT \quad (3-3)$$

$$= 3kT \ln \left(\frac{N_A(x)}{10^{17}} \right) \quad (3-4)$$

where the second expression may be used for the regions with high doping above 10^{17} cm^{-3} . The total bandgap reduction $\Delta E_g(x)$ is the sum of the two terms in (3-3) and (3-4), and the effective bandgap $E_g(x)$ is modeled as

$$\begin{aligned} E_g(x) &= E_g(0) + [E_g(W_{\text{BAG}}) - E_g(0)] \frac{x}{W_{\text{BAG}}} \\ &= E_g(0) - \Delta E_g \frac{x}{W_{\text{BAG}}} \end{aligned} \quad (3-5)$$

where $E_g(0) = E_{g, \text{Ga}} - \Delta E_g(0)$ and $E_g(W_{\text{BAG}}) = E_{g, \text{Ga}} - \Delta E_g(W_{\text{BAG}})$ with $E_{g, \text{Ga}}$ is the bandgap of intrinsic GaAs. Bandgap narrowing is due to heavy doping effects and due to Ga are then accounted for in a unified way.

The permittivity of the GaIn alloy layer can be different from that of Ga and will be a function of Ga mole fraction x . However, since this material property has not been reported so far, we characterize it reasonably using a linear interpolation between the permittivities of Ga and In as a function of Ga mole fraction x

$$\epsilon_{\text{GaIn}}(x) = \epsilon_{\text{In}} + (\epsilon_{\text{Ga}} - \epsilon_{\text{In}}) \cdot x \quad (3-6)$$

where ϵ_{Ga} and ϵ_{In} are the permittivities of Ga and In, respectively. This approximation should give better explanation of the quantum space-charge dynamics associated with the moving boundaries, which are functions of the

junction bases, than the simple junction capacitances based on a constant permittivity.

The *epi-collector* doping density N_{epi} is assumed to be constant, even though some technologies use non-uniform profiles to effect a trade-off between the collector junction capacitance and its breakdown voltage [Uyot01]. This enables use of the basic collector analysis described in Chapter 2. A conventional emitter structure, with a polysilicon contact is assumed, and the carrier transport is modeled based on [Foot01].

3.3.3 Electrostructural Ring Analysis

We reformulate the MMSPICE-1.0BET base analysis for the transport (injection) current and excess carrier charge (Graf5) to account for heterostructures as in SiGe-base BICs as well as Si BICs. A typical (one-dimensional) Ge mole fraction profile and an intrinsic-region impurity-density profile are depicted in Fig. 3-1. Bandgap profile under the thermal equilibrium across the intrinsic device is also illustrated in Fig. 3-2. As illustrated in Fig. 3-3, the conduction band edge is not necessarily parallel to the valence band edge due to the bandgap reduction in the base region, and the effective electric field acting on the (intrinsic) electrons is generally different from that acting on the (majority) holes, which is not the case in homogeneous devices. Thus we need to derive new transport current and base transit time expressions, which reflect the bandgap variation in the base under the built-in fields [Koo05].

The built-in field E_b acting on the minority electrons can be derived with the negligible hole current in the base and with nondegenerate carrier densities assumed. Noting that $E_{\text{eq}} = E_n + E_p$, we have $qE_n = dE_p/dx = -qE_b + dE_p/dx$, where the effective field E_b on the holes is defined by $E_b = E$ (or

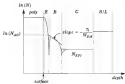


Fig. 3.1 Assumed one-dimensional carrier structure, (a) doping, and (b) Ge mole fraction profile

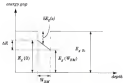


Fig. 3-3: Energy band gap diagram for the DBIT structure in Fig. 3-1. ΔE reflects all the bandgap reductions from both Ge concentrations and high doping effect.

detected balance). Then E_x is expressed in terms of base-region parameters (see Appendix A):

$$E_x = -\frac{kT}{qA \exp\left(-\frac{qV}{kT}\right)} \frac{d}{dx} \left[p \exp\left(-\frac{E_g}{kT}\right) \right] \quad (3-7)$$

where p denotes the hole concentration. In the derivation of (3-7), it is assumed that Einstein relation can be applied without significant error at doping level as high as 10^{19} cm^{-3} . Combining this result with the electron-current equation yields a collector-current expression

$$\begin{aligned} J_{cx} &= qv_{ex}nE_x + qD_n \frac{dn}{dx} \\ &= qv_{ex} \left(\frac{dn}{dx} + \frac{n}{p} \frac{dp}{dx} + \frac{n}{kT} \frac{dE_g}{dx} \right) \\ &= qv_{ex} \left[\frac{dn + N_A \frac{dn}{dx}}{n + N_A} + \frac{n}{n + N_A} \frac{dN_A}{dx} + \frac{n}{kT} \frac{dE_g}{dx} \right] \end{aligned} \quad (3-8)$$

A. Low-Doping Case: For low injection, $N_A \gg n$ for all x , and (3-8) can be simplified to

$$\begin{aligned} J_{cx} &\approx \frac{qv_{ex}}{N_A} \left[\frac{d}{dx} (nN_A) + \frac{nN_A}{kT} \frac{dE_g}{dx} \right] \\ &= \frac{qD_n}{N_A} \left[\frac{d}{dx} (nN_A) + \frac{nN_A}{kT} \frac{dE_g}{dx} \right] \\ &= \frac{qD_n}{N_A \exp\left(-\frac{qV}{kT}\right)} \frac{d}{dx} \left[nN_A \exp\left(-\frac{qV}{kT}\right) \right] \end{aligned} \quad (3-9)$$

where $\eta' = (q + kT/qT) / W_{BB}$. In deriving (3-9), we have used (3-1) and (3-6). Integrating (3-9) across the QSE yields

$$J_{\text{tot}} = qD_{\text{tot}}W' \frac{n(x_0) \exp(\eta'W_0) - n(x_0)}{\exp(\eta'W_0) - 1} \quad (3-10)$$

where $W_0 (= x_0 - x_d)$ represents the QFB width modulated by the applied positive bias, and D_{tot} is the averaged diffusivity defined by

$$D_{\text{tot}} = \int_{x_0}^{x_d} N_A(x) \exp\left(-\frac{q\phi x}{kT W_{\text{tot}}}\right) dx / \int_{x_0}^{x_d} \left(\frac{N_A(x)}{D_n(W_0)}\right) \exp\left(-\frac{q\phi x}{kT W_{\text{tot}}}\right) dx. \quad (3-11)$$

We can decompose (3-11) into two terms, which are associated with the emitter and collector junctions, respectively

$$J_{\text{tot}} = J_{\text{emf}} + J_{\text{colf}} \quad (3-12)$$

where

$$J_{\text{emf}} = qD_{\text{em}}W' \frac{\exp(\eta'W_0)}{\exp(\eta'W_0) - 1} n(x_0) \quad (3-13)$$

and

$$J_{\text{colf}} = qD_{\text{col}}W' \frac{n(x_0)}{\exp(\eta'W_0) - 1} \quad (3-14)$$

Integrating (3-5) from x_0 to an arbitrary position x within the QFB, we can describe the excess minority-carrier density, $n(x)$, explicitly. The integration of the injected excess minority-carrier density across the QFB then yields a total charge Q_{tot} , which can be expressed as the sum of two terms

$$Q_{\text{tot}} = Q_{\text{emf}} + Q_{\text{colf}} \quad (3-15)$$

where Q_{emf} and Q_{colf} are associated with the emitter and collector junctions, respectively, and are given by

$$Q_{\text{emf}} = \frac{qN_A}{\eta} \left[\frac{\eta'W_0 \exp(\eta'W_0)}{\exp(\eta'W_0) - 1} - 1 \right] n(x_0) \quad (3-16)$$

and

$$Q_{\text{coll}} = \frac{qA}{\eta} \left[1 - \frac{\eta^* W_d}{\exp(\eta^* W_d) - 1} \right] n(x_d) \quad (3-17)$$

B. High-Injection Case: With a similar procedure as in the low injection case, the expressions for the collector current and the total excess-carrier charge for high injection can be derived. For $N_A \ll n$ assumed for all x in the QNR region, (3-6) simplifies to

$$\begin{aligned} J_{\text{coll}} &= qnD_n \left[\frac{dn}{dx} + \frac{n}{2kT} \frac{dE}{dx} \right] \\ &= \frac{2qD_n}{\exp(-\eta^* x) - 1} \frac{d}{dx} [\exp(-\eta^* x)] \end{aligned} \quad (3-18)$$

where $\eta^* = \Delta E / 2kT W_d$. Integrating (3-18) by parts across the QNR, we get a collector-current-expression for high injection given by

$$J_{\text{coll}} = J_{\text{coll}}^* - J_{\text{coll}} \quad (3-19)$$

where

$$J_{\text{coll}}^* = qnD_n \eta^* \frac{\exp(\eta^* W_d)}{\exp(\eta^* W_d) - 1} n(x_d) \quad (3-20)$$

and

$$J_{\text{coll}} = qnD_n \eta^* \frac{n(x_d)}{\exp(\eta^* W_d) - 1} \quad (3-21)$$

The averaged diffusivity $\bar{D}_{\text{coll}}$ for high injection, is defined as

$$\bar{D}_{\text{coll}} = \int_{x_1}^{x_2} \exp(-\eta^* x) dx \int_{x_1}^{x_2} \frac{\exp(-\eta^* x)}{D_n (W_d)} dx \quad (3-22)$$

The total injected excess carrier charge for the high injection is expressed in the same manner as in the low injection case

$$Q_{\text{ex}} = Q_{\text{exL}} + Q_{\text{exH}} \quad (3-25)$$

where

$$Q_{\text{exH}} = \frac{qA}{v} \left[\frac{\eta'' W_p \exp(\eta'' W_p)}{\exp(\eta'' W_p) - 1} - 1 \right] n(x_p) \quad (3-26)$$

and

$$Q_{\text{exL}} = \frac{qA}{v} \left[1 - \frac{\eta'' W_p}{\exp(\eta'' W_p) - 1} \right] n(x_p) \quad (3-27)$$

C. General Equation. Now we need to generalize the expressions for the collector current and the total excess carrier charge in the QNS to account for all injection levels. We do this, following (Joshi), properly accounting for the moving QNS boundary conditions, but simply using linear interpolation between both extremes with $n(x_p)$ and $n(x_c)$ as weighting parameters. The generalized expressions of this collector and total excess-carrier charge are then given as

$$J_c = \frac{J_{\text{exL}} \frac{N_A(x_p) + J_{\text{exH}}/n(x_p)}{N_A(x_p) + n(x_p)} + J_{\text{exH}} \frac{n(x_p)}{N_A(x_p) + n(x_p)}}{\frac{N_A(x_p) + J_{\text{exH}}/n(x_p)}{N_A(x_p) + n(x_p)} + \frac{J_{\text{exL}} \frac{N_A(x_p) + J_{\text{exH}}/n(x_p)}{N_A(x_p) + n(x_p)} + J_{\text{exH}} \frac{n(x_p)}{N_A(x_p) + n(x_p)}}} \quad (3-28)$$

and

$$Q_{\text{ex}} = \frac{Q_{\text{exL}} \frac{N_A(x_p) + Q_{\text{exH}}/n(x_p)}{N_A(x_p) + n(x_p)} + Q_{\text{exH}} \frac{n(x_p)}{N_A(x_p) + n(x_p)}}{\frac{N_A(x_p) + Q_{\text{exH}}/n(x_p)}{N_A(x_p) + n(x_p)} + \frac{Q_{\text{exL}} \frac{N_A(x_p) + Q_{\text{exH}}/n(x_p)}{N_A(x_p) + n(x_p)} + Q_{\text{exH}} \frac{n(x_p)}{N_A(x_p) + n(x_p)}}} \quad (3-29)$$

Both (3-28) and (3-29) need boundary conditions at the moving edges of the QNS region to get explicit values. The required values on the boundary conditions are the locations of the moving boundaries (x_p and x_c) and the excess

carrier densities ($n(x_E)$ and $n(x_C)$). These are obtained through the collector and emitter region analyses and will be described in the subsequent sections. However, we develop a model for the energy barrier formation first since it affects the collector side boundary conditions.

3.3 Energy Barrier Formation

IIST devices have unique characteristics related to the bandgap discontinuity at the base-collector heterojunction, which requires the extension of the HSPICE-1 SPBT collector analysis in Chapter 2. A model of the potential energy barrier formation for small modulations is developed in this section. This will be linked to the boundary conditions discussed in the subsequent section 3.4.

The effective QNR region extends beyond the metallurgical collector junction as the current becomes greater than the base push-out onset current. In conventional transparent IHTs, which can have an abrupt bandgap variation through the bulk, the onset current is closely defined as noted in Chapter 2 since the subsequent base push-out is accounted for by injected holes, which are not prohibited by any valence-band barrier.

The onset of the base push-out in the IIST, however, can occur at a higher current because of the bandgap discontinuity at the collector junction. At the current level where the junction SCH collapses completely, the bandgap discontinuity appears in the valence band, which acts as an additional barrier for holes and thus the hole injection is suppressed exponentially. The lack of holes to neutralize the electrons in the vicinity of the junction suppresses the base push-out. Thus the base push-out does not occur immediately but is deferred to a higher current [Nish].

We develop a physical model for the energy barrier which is applicable in most situations by extending the insight afforded by Chapter 3. We begin with the junction SCL width W_{SCL} , whose expression is derived in Chapter 3, but quoted here for the convenience of our discussion.

$$W_{\text{SCL}} = \left[\frac{2\epsilon_0\epsilon_{\text{Si}}(V_{\text{C}} - V_{\text{BC}} - I_{\text{C}}R_{\text{BT}}/q)}{I_{\text{C}} - I_{\text{C}}'} \right]^{\frac{1}{2}} \quad (3-28)$$

Taking derivatives of (3-24), we find that W_{SCL} increases with collector current for $V_{\text{BC}} < V_{\text{BCRT}} = V_{\text{C}} - I_{\text{C}}R_{\text{BT}}$, which leads to the non-chase quasi-saturation, and decreases otherwise, which results in the chase quasi-saturation. We develop our model for each case.

A. Non-Chase Quasi-Saturation. The expression of W_{SCL} eventually ends with the entire gap layer depleted with the space charge, which is partially compensated with the mobile carriers. The collector current when $W_{\text{SCL}} = W_{\text{BT}}$ is defined as $I_{\text{ACC/BT}}$ (see Chapter 2):

$$I_{\text{ACC/BT}} = I_{\text{C}} + \frac{1}{R_{\text{BT}} - R_{\text{BT}}'}(V_{\text{BC}} - V_{\text{BCRT}}) \quad (3-29)$$

Further increase of the current beyond $I_{\text{ACC/BT}}$ continues to incorporate the depleted space charge eventually changing the polarity of the space charge from the positive to the negative. Until I_{C} reaches $I_{\text{ACC/BT}}$, which is the onset of the operational base push-out, there remains still a depleted junction (SCL) in the base region. This SCL collapses completely when the collector current reaches to $I_{\text{ACC/BT}}$ (see Chapter 2):

$$I_{\text{ACC/BT}} = I_{\text{C}} + \frac{V_{\text{BC}} - V_{\text{BCRT}}}{R_{\text{BT}}} \quad (3-30)$$

The schematic band diagrams of $\beta_{\text{ACR-type}}$ are shown in Fig. 3-3 for the heterojunction and homojunction devices. Note that the built-in potentials are not equal to each other because of the difference in base materials although both emitter dopant profiles may be identical. However, the junction behavior in the case for both devices until the critical current is reached. The presence of bandgap discontinuity at the collector junction makes the difference beyond $\beta_{\text{ACR-type}}$.

A qualitative but physical accounting for the barrier formation and its effects can be made as follows. The bandgap discontinuity appears in the collector band and forms an energy barrier to the flow of the holes from the base. The resistance to the passage of hole movement at the junction delays the base push out until the required holes for the quasi-neutrality is established. At the same condition, the emitter n_p layer is space charged by means of the excess electrons supporting the current. Further increase of the current can only be sustained by the increased amount of the electron carriers since carrier velocity is almost saturated. The accumulation of the (negative) charge in the collector side of the junction induces a (positive) charge layer in the base side to preserve the total charge neutrality through the base and collector regions. Conceptually, the thickness of this layer will be very small since the carrier density of the holes is very high compared to that of the electrons and thus this space charge layer can be regarded as consisting of a dipole layer. The electric field originating from this space charge is against the flow of the electrons, which means equivalently a potential barrier for the electrons from the junction. This situation depicts a negative feedback since the tendency of the current increase is suppressed by the induced potential barrier. The potential barrier effectively blocks the electrons injected from the emitter junction and as a result electrons accumulated in the quasi-neutral base region. This in turn

induces more electrons from the emitter to maintain the increased current, which results in the rapid increase of the emitter junction bias. The base current also increases rapidly to supply holes for the charge neutralization in the base.

While the potential barrier for the electrons exists in the conduction band at the collector junction, the entire *epi-collector* layer remains space charged. A consistent view of the relation between the energy barrier band diagram and the electric field in the *epi-collector* is illustrated in Fig. 3-4. We assume here that the potential drop across the dipole layer is negligible because of its small thickness. Then, as the current increases the bands in the *epi-collector* shift upward, effectively raising the valence-band barrier, which is effectively supported entirely by the dipole layer. Knowing that the integral of the electric field (which is obtained as a solution of Poisson's equation) across the junction and *epi-collector* is equal to $\phi_C = V_{BC}$, and assuming that the magnitude of the potential barrier at the junction is Φ , we can express the potential drop across the *epi-collector* region as $\phi_E = V_{BC} - \Phi$. Now we derive a model for the magnitude of Φ using insight from Chapter 2. To simplify the model development, we assume the base push-out happens after the bandgap difference $\Delta E_g/(W_{AB})$ at the collector junction appears fully as the energy barrier in the conduction band. The corresponding current and potential barrier are denoted as I_E and $\Phi_E (= \Delta E_g/(W_{AB})/q)$. Since the carrier velocity is assumed to be saturated within the RCE, the maximum electric field at the junction is pinned to the threshold field E_g . The subsequent increase of the potential barrier with the collector current is reflected by the increase of the slope of the electric field, which is proportional to the current, as illustrated in Fig. 3-5. Solving Poisson's equation by integrating this field yields

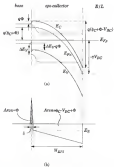


Fig. 2-6 (a) Ideal diagram of a hetero-collector junction for the case of a collector band further separated. (b) A corresponding electric field picture showing the existence of a dipole layer as the result of the band separation.

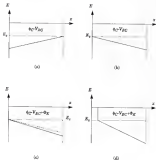


Fig. 3-4. Electric field patterns for the various current levels under the conditions that V_{acc} is $V_{acc,cr}$: (a) $I_0 < I_{acc,cr}$, (b) $I_0 = I_{acc,cr}$, (c) $I_0 > I_{acc,cr}$, (d) $I_0 > I_{acc,cr}$. (c) $I_0 = I_{acc}$, and (d) $I_0 > I_{acc}$.

$$V_C - V_{BE} + \Phi = -\frac{qW_{\text{HT}}^2}{2\epsilon} \Delta n + E_2 W_{\text{HT}} \quad (3-31)$$

where Δn is the excess electron carrier density, which can be expressed in terms of I_C

$$q \left(\frac{I_C}{qAa_f} - N_{\text{HT}} \right) = q \Delta n. \quad (3-32)$$

Substituting (3-32) into (3-31), an expression for Φ is obtained

$$\Phi = -\frac{qW_{\text{HT}}^2}{2\epsilon} \left(\frac{I_C}{qAa_f} - N_{\text{HT}} \right) + E_2 W_{\text{HT}} - (V_C - V_{BE}). \quad (3-33)$$

According to (3-33), the energy barrier grows linearly with the increase of the collector current. This energy barrier builds up until the quasi-neutrality is established near the junction by reducing the electron flow from the emitter and increasing the hole flow from the base by lifting up the valence band. The linear relation between I_C and Φ is important to understand the abrupt increase of the base current and the new flat characteristic in the forward plot. This subject will be discussed later in detail.

B. Ohmic Quasi-Saturation. According to (3-27), W_{HTC} shrinks as the current increases with V_{BE} fixed, and eventually vanishes at the current I_{HTC} (see Chapter 5):

$$I_{\text{HTC}} = I_C = \frac{V_{\text{BE}} - V_{\text{BET}}}{R_{\text{HT}}} = \frac{I_C - V_{\text{BE}}}{R_{\text{HT}}}. \quad (3-34)$$

Again, the junction behavior is same for both homogeneous and heterojunction devices until this current is reached. With the same physical reason as in the non-sharp quasi-saturation, the base push-out cannot occur if a bandgap discontinuity exists at the collector junction of the HBT. In contrast to the former case, since the field intensity is below the threshold field E_p , recomb

electrons are not necessary to support the increased current, the current is supported by the carrier velocity increasing in proportion to the field. Without the base widening, the increase of the electric field builds up a potential barrier in the conduction band at the junction to reduce the electron flux into the collector. Thus the holes are injected more as a result of the barrier reflection in the valence band. This action continues until the quasi-neutrality is established. The expression of the energy barrier ϕ_b in this case, however, is different from (3-33) since there are no excess carriers in the *np* layer. We can suppose that there can be two different situations during the buildup of the energy barrier with the current increase. The first case is that until the field intensity reaches the E_g , the quasi-neutrality near the junction is not achieved, and thus the entire *np* layer is space-charged to support the increased current, and at a higher current the base pinch-out occurs in the same way as described in the one-dimensional situation case. The second case is that the quasi-neutral condition is met when the electric field intensity is less than E_g , whereafter the base region extends into the collector as in conventional devices. The variations of the schematic electric field patterns in the *np* collector with the current increase are shown in Figs. 3-4 and 3-5 for both cases. We derive two corresponding expressions for Φ below.

We consider the case depicted in Fig. 3-4 first. Note that the electric field is uniform with the field intensity of $I_C/(\mu_p q N_A W)$, which is less than E_g as long as the current level is below I_g . Since the total area obtained by integrating the field across the *np*-collector is equal to $\phi_C - V_{BC} + \Phi$, the magnitude of Φ , until I_C reaches to I_g , can be given by

$$\Phi = I_C R_{g21} - (\phi_C - V_{BC}) \quad (3-34)$$

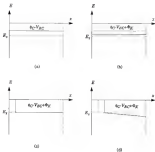


Fig. 3-4 Electric field profiles for the vacuum-current levels under the conditions that $V_{acc} < V_{acc} < \Phi_E$ and $I_0 < I_{cr}$. (a) $I_0 = I_{acc}$, (b) $I_0 = I_{cr}$, (c) $I_0 = I_{cr}$, (d) $I_0 > I_{cr}$.

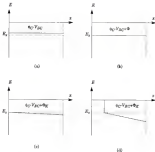


Fig. 5-1 Electric field patterns for the various current levels under the condition that $V_{GC} < V_{GC} \leq v_C$ and $I_G \leq I_G$ (a) $I_G = I_{GC}$, (b) $I_G = I_G$, (c) $I_G = I_G$, (d) $I_G = I_G$.

The device base push-out continues until I_B reaches I_{B_0} , and then non-linear base push-out occurs as illustrated in Figs. 3-8(i) and 3-8(j).

In the second case depicted in Fig. 3-7, the maximum energy barrier $q\Phi_B$ is not reached at the current I_{B_0} yet and thus I_B is greater than I_{B_0} as shown in Fig. 3-10. Thus the dependence of the Φ for $I_B < I_{B_0}$ is the same as that in (3-38). For $I_B > I_{B_0}$, the base push-out will not happen until the full bandgap difference appears in the conduction band as in Fig. 3-9(i). An SCR identical to that in Fig. 3-8(i) is induced by the excess electrons to support the increased current, therefore we can use the same equations as that in (3-35).

3.4 Boundary Conditions

The base region transport current and excess carrier charge expressions developed in section 3.2 are linked to the terminal biases through the emitter-base and collector-base boundary conditions, which are discussed in this section. For a homogeneous BJT, the required boundary value expressions are developed and described in Chapter 2 comprehensively based on physical insight. We extend the expressions to make it applicable to the SiGe-base HBT model as well as the BJT model. The underlying formulation is identical in both devices until the junction SCE collapses completely, i.e., the current level reaches the conventional limit of the base push-out, except of course for the accounting of the bandgap variation in the base region. The energy barrier formation described in the previous section makes the fundamental difference between the HBT and BJT with respect to the collector junction behavior. Since the energy barrier to the electron flow leads to steep increases of excess electrons in the QNR (which cannot be seen in the BJT), and its formation is sensitive to high current operation, the accounting of this new effect is essential for accurate circuit simulations, and we implement it through the

collector-side boundary conditions as described in this section. The emitter-side boundary values are also modified due to the effect of the bandgap narrowing and we describe it here.

3.6.1 Emitter-Side Boundary Values

In a highly doped device with high doping concentrations and shallow emitter junctions, the voltage drop across the bulk emitter is negligible even at the high operating currents. Thus the effective junction bias V_{BE} can be approximated as the forward base-emitter bias minus the voltage drop at the emitter contact and base resistance. The emitter junction (EJ) width is a function of V_{BE} . The moving boundary location x_E in the base (which is represented as WBE in our model circuit) is modeled as $w_E[baseC, baseR]$

$$x_E = WBE - WBE0 \left[\frac{FE - V_{BE}}{FE} \right]^{ME} \quad \text{for } V_{BE} < FE - \Delta\phi_E \quad (3-36)$$

$$= WBE0 \exp \left(- \frac{V_{BE}}{\Delta V_T} \right) \quad \text{for } V_{BE} > FE - \Delta\phi_E \quad (3-37)$$

where FE is the built-in potential across the emitter junction and is evaluated by solving the one dimensional Poisson's equation based on the doping profile, reflecting the bandgap and permittivity variation in the base. $WBE0$ denotes the width of the base-side depletion region (x_E) of the emitter junction (EJ) at thermal equilibrium. ME explains the variation of WBE when the position bias changes; $WBE0$ and $\Delta\phi_E$ are obtained with the continuity mesh form of WBE and its derivative at $V_{BE} = FE - \Delta\phi_E$ with respect to V_{BE} .

$$WBE0 = WBE0 \left[\frac{\Delta V_T \cdot ME}{FE} \right]^{ME} \exp \left[\frac{FE - \Delta V_T \cdot ME}{\Delta V_T} \right] \quad (3-38)$$

$$\Delta\phi_E = \Delta V_T \cdot ME \quad (3-39)$$

Applying quasi-equilibrium and quasi-neutrality at x_D , the μ_n product can be expressed as

$$\mu_n = [N_A(x_D) + n(x_D)](N_A(x_D) + n_{\text{eq}}^D(x_D) \exp(-\frac{qV_{BD}}{kT})) \quad (3-40)$$

where the effective intrinsic carrier density $n_{\text{eq}}^D(x_D)$ is position dependent and given by

$$n_{\text{eq}}^D(x_D) = n_i^D \exp(-\frac{E_{\text{eff}}(x_D) - E_F(x_D)}{kT}) \quad (3-41)$$

From (3-40), the boundary condition $n(x_D)$ for the base transport is given explicitly in terms of V_{BD} , $E_F(x_D)$, and $N_A(x_D)$.

3.4.3 Collector-Side Boundary Values

The moving boundary x_C and the maximum minority carrier density $n(x_C)$ at the collector-side of the QTE reflect the complicated collector junction behavior including the base push-out (Kirk effect). In 3D-QFTs, the bandgap variation in the base region, implying the energy barrier formation, as described in section 3.3, leads to the more complicated boundary value expression. We describe them for three operational modes below.

A. Junction-SCH Mode In this operational mode, the junction SCH exists and thus the boundary x_C is moving depending on the bias and current. The formulation of the boundary values in this mode is same to that developed for the heterojunction BJT in Chapter 2. The quasi-neutral edge x_C is located by equating the total space charge in the qn region to the junction SCH in the base region. From (3-38) and (3-39), and knowing that the base doping profile is exponentially represented by two parameters N_{BD} and γ , x_C is expressed as

$$n_C = N_{\text{bar}} - N_{\text{BC}} = -W_{\text{bar}} - \frac{W_{\text{bar}}}{q} \ln \left[\frac{(U_C - U_C^0) W_{\text{bar}}/q}{k_B T_{\text{bar}} (W_{\text{bar}}/q - U_C^0) W_{\text{bar}}} + 1 \right] \quad (2-42)$$

To define n at (n_C) , the QPL equation $V_{\text{BC},C}$ at n_C is modeled in the same way as in Chapter 3 (see section 3.1.3). Applying quasi-neutrality and quasi-equilibrium, we get

$$[P_A(n_C) + n(n_C)] + (n_C) = n_a^0(n_C) \exp \left(-\frac{V_{\text{BC},C}}{k_B T} \right) \quad (2-43)$$

where $n_a(n_C)$ is the effective intrinsic carrier density reflecting the bandgap variation at n_C . (2-43) defines $n(n_C)$ in terms of $V_{\text{BC},C}$, $E_g(n_C)$, and $N_A(n_C)$.

B. Energy Barrier Formation Mode. This operational mode occurs after the positive dcCB collapse and until the base peak-out starts as a result of the maximum barrier formation. The boundary between n_C is fixed at W_{bar} and the remaining problem is to solve $n(W_{\text{bar}})$. Depending on the bias condition, three ways of energy barrier formation are possible as described in section 3.3. The energy barrier formation results in the increase of $n(n_C)$ since the QPL equation at the junction is enhanced further by the increase of the potential barrier. Applying quasi-neutrality and quasi-equilibrium at the junction leads to the following relation:

$$[N_A(W_{\text{bar}}) + n(W_{\text{bar}})] + n(W_{\text{bar}}) = n_a^0(W_{\text{bar}}) \exp \left(\frac{U_C + \Phi}{k_B T} \right) \quad (2-44)$$

where $n_a(W_{\text{bar}})$ is the effective intrinsic carrier density at W_{bar} . This relationship is illustrated through the energy band diagram shown in Fig. 3.8.

C. Base Peak-Out Mode. Further increase of the collector current eventually induces the base peak-out. Then the op-collector consists of two different regions according to their electrical characteristics, a quasi-neutral region (NQD), and a non-linear current-induced SCR or an almost unmodulated

(uniform field) region depending on the bias condition. If the hole peak out-curve under the non-quasi-saturation bias condition, the field adaptation across the current-induced SCR is equal to the potential $\phi_C - V_{BC} + \phi_D$, and the SCR width W_{SCR} is expressed as

$$W_{SCR} = \frac{\text{Im}_A (\phi_C - V_{BC} + \phi_D)}{eR_{sh}^2 + \sqrt{(\pi R_{sh}^2)^2 + \frac{\text{Im}_A^2}{A} (\bar{L}_B - \bar{L}_D) (\phi_C - V_{BC} + \phi_D)}} \quad (3-42)$$

where the carrier velocity is assumed to be saturated. Otherwise, if the hole peak out-curve occurs under the ohmic quasi-saturation bias condition, since the potential drop across the ohmic region is equal $\phi_C - V_{BC} + \phi_D$, the width of this unsaturated region is expressed as

$$W_{unsat} = (\phi_C - V_{BC} + \phi_D) \frac{qA_s \mu_{eff}}{I_C} \quad (3-43)$$

The existence of the QNR in the op-collector for this case complicates the boundary value expressions since we do not have explicitly the QPL equations at the collector junction. To solve this problem, the I_C expression in terms of QPL equations at both edges of the op-QNR, derived by Rail et al. [Rail05], has been used, since the QPL equation at the far edge of QNR from the junction is explicitly described (see section 3.3.2). If $\bar{L}_D$ is known, the QPL equation at the junction can be calculated according to the I_C expression. We extend the expressions in (3-43) to make it applicable to the full EIT as well as the EIT with no bandgap discontinuity at the collector junction. Assuming the hole current is negligible within the QNR is ϵ , the hole quasi-Fermi level is constant, we express the collector (electron) current as

$$I_C = -qA_s \mu_{eff}^n \frac{dn}{dx} \quad (3-44)$$

where qp is the QFL separation for electrons (QFL) within the region. Assuming negligible recombination within the QNL, and integrating (3-47) by parts across the QNL, we can have

$$I_C = -\frac{qA\phi_0}{W_{QNL}} \int_0^{x(x_{end})} n(x) dx \quad (3-48)$$

where the origin of x is at the collector junction. The qp product is related to the QFL separation, and is expressed as

$$p(x)n(x) = n_0^2 \exp\left(\frac{q\phi}{kT}\right) \quad (3-49)$$

Applying quasi-neutrality to (3-49), we describe $n(x)$ in terms of ϕ . Substituting it into the integral in (3-48) yields

$$\begin{aligned} \int_0^{x(x_{end})} n(x) dx &= -N_T \{ \phi(x_{end}) - n(W_{QNL}) - N_{DNL} \ln \left[\frac{n(x)}{n(W_{QNL})} \right] \} \\ &= \frac{N_{DNL}}{q} (V_{BCL} - V_{BCL}) \quad \} \end{aligned} \quad (3-50)$$

where $V_{BCL} = \phi(0)$ and $V_{BCL} = \phi(W_{BCL})$. In the homogeneous QFL, we showed in Chapter 2 that the QFL separation V_{BCL} can be modeled as ϕ_C . In the heterojunction QFL, with the same physical reasoning, this is modeled as

$$V_{BCL} = \phi_C + \phi_E \quad (3-51)$$

which refers to ϕ_C for the homogeneous device with no energy barrier formation. Substituting (3-48), (3-49), and (3-50) into (3-45), we finally express the collector current as terms of the physical parameters pertaining to the quasi-neutral opt-collector region as

$$I_E = -\frac{qA_E V_T N_{BT}}{V_{GBT}} \left[(E(V_{BGT}) - E(V_{BGT})) - \ln \left(\frac{1 + E(V_{BGT})}{1 + E(V_{BGT})} \right) + \frac{V_{BGT} - V_{BGT}}{V_T} \right], \quad (3-52)$$

where

$$E(V) = \sqrt{1 + \left(\frac{V_0}{V_{GBT}} \right)^2} \exp \left(\frac{V}{V_T} \right). \quad (3-53)$$

The equation (3-52) relates V_{BGT} to I_E , i.e., with I_E is known, then V_{BGT} can be calculated. After V_{BGT} is obtained, the excess carrier density $n(V_{BGT})$ is calculated by using

$$n(V_{BGT}) = n(V_{BGT}) + N_A(V_{BGT}) = n_0(V_{BGT}) \exp \left(\frac{V_{BGT}}{V_T} \right) \quad (3-54)$$

where the effective intrinsic carrier density $n_0(V_{BGT})$ reflects the bandgap discontinuity at the junction.

3.3 Complete Model Implementation

The network representation of the charge-based 3D1-layer HBT model is shown in Fig. 3-5. The implicit characterization of I_E is done through iterative solution of base and collector analysis described previously in this chapter. The submicron-gap collector current I_{EC} through the peripheral (npe) base-collector junction is characterized physically as described in [Choi03]. Regarding I_B and charging currents, we reuse the previous implementation and describe needed updates.

The characterization of I_B in Fig. 3-5 reduces to post-processing after the main iterative solution is attained. The dc base current, excluding the input resistance component discussed in Chapter 4, is modified as [modified]

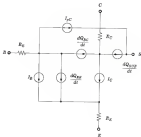


Fig. 3-4: Network representation of charge-based SOI-bias JFET in triangular mode

$$\begin{aligned}
I_d = & A_d \cdot JEO \cdot \left[\exp\left(\frac{V_{dR}}{V_T}\right) - 1 \right] + A_E \cdot JEO \cdot \left[\exp\left(\frac{V_{dR}}{JEEB \cdot V_T}\right) - 1 \right] \\
& + \frac{Q_{TE}}{TE} + \frac{Q_{TC}}{TC} \\
& + 1 \cdot JEO \cdot (WE + LE) \cdot \left[\exp\left(\frac{V_{dR}}{JEEB \cdot V_T}\right) - 1 \right]
\end{aligned} \quad (3-50)$$

where JEO , $JEEB$, $JEEA$, TE , TC , $JEOB$, and $JEEB$ are model parameters, and WE and LE are the effective width and length of the emitter. The first term in (3-50) represents the back-injection emitter current component, for which we use the same formulation as (Jeeffix) since the existence of the bandgap discontinuity at the emitter junction does not change the hole injection. JEO is indeed a constant, albeit dependent on the emitter morphology. The second term describes the EBE recombination current in the (integrated) emitter-base junction EBJ . The third and fourth terms model the recombination currents in the QNB and quasi-neutral base extensions in the opt-collector when base perfect occurs. The last term models the EBE recombination current in the (spherical) emitter-base junction EBJ (Jeeffix).

Tracking the currents of the regional charges and partitioning them physically, we can effectively characterize transistors/rectifiers and their some non-quasi static effects. Calculating the hole charges injected into each region from the base terminal and applying quasi neutrality give a consistent description of the regional charges as described below.

The total hole charge in the QNB can, based on quasi neutrality, be expressed in terms of electron and acceptor dopant charge (Jeeffix):

$$\begin{aligned}
Q_{QNB} = & qA \int_{x_0}^{x_1} n_0 dx + qA \int_{x_0}^{x_1} N_A dx \\
= & qA \int_{x_0}^{x_1} n_0 dx + qA \int_0^{x_1} N_A dx - qA \int_0^{x_0} N_A dx - qA \int_{x_1}^{x_2} N_A dx
\end{aligned}$$

$$= Q_N + Q_{DE} - Q_{SE} - Q_{SC} \quad (3-34)$$

where, from the exponential doping profile assumed, we have

$$Q_{DE} = \frac{qAN_{DE}W_{DE}}{\eta} \left[1 - \exp\left(-\frac{W_{DE}}{W_{DE}}\right) \right] \quad (3-37)$$

$$Q_{SE} = \frac{qAN_{SE}W_{SE}}{\eta} \left[\exp\left(-\frac{W_{SE}}{W_{SE}}\right) - \exp(-\eta) \right] \quad (3-38)$$

with $Q_{SD} = qAN_{SD}W_{SD} [1 - \exp(-\eta)] / \eta$

Physically, Q_{DE} , expressed in (3-37), is modulated by both emitter and collector junctions in transient operation, and thus it is more physical to partition the charge into two components. The partitioning has been implemented by introducing an empirical partitioning factor f_D as in [Fau08, Aust04]

$$Q_N = Q_N^E + Q_N^C = Q_{DE}f_D + Q_{SE}(1 - f_D) \quad (3-39)$$

where f_D is assigned a value between zero and one

The depletion charge Q_{DE} within the emitter junction-SCR is dependent on N_{DE} only, and the associated charging current flows through the emitter junction only. The depletion charge Q_{SE} in the collector junction-SCR, however, is modulated by both junction bases since the collector current affects the SCR and is dependent on the emitter junction base. Furthermore, Q_{SE} defines charging current components associated with the emitter as well as the collector terminals. To define the needed partitioning, we refer to Chapter 3 and relate Q_{SC} to the n -collector regional charges

$$-Q_{SC} = Q_{SE1} + Q_{SE2} \quad (3-40)$$

where

$$Q_{SE1} = qAN_{SC} \left(N_{DE1} - \frac{J_D}{qA J_D} \right) = \frac{W_{SC1} J_D}{v_D} - \frac{W_{SC1} J_D}{v_D} \quad (3-41)$$

and

$$Q_{BE} = 0 \quad (3-52)$$

in the unbalanced forward active mode, or

$$Q_{BE} = qA W_{BC} \left(R_{BE} - \frac{I_C}{qA n_{BE}} \right) = \frac{W_{BE}}{v_E} I_E - \frac{W_{BE}}{v_E} I_C \quad (3-53)$$

and

$$Q_{BC} = \frac{qA}{W_{BE}} (R_C - V_{BC}) = \frac{W_{BE}}{W_B} (I_C - I_E) = qA F_B \quad (3-54)$$

in the unbalanced forward active mode where the entire n_p -collector is space-charged. If the collector junction BCN collapses, the sum of Q_{BE} and Q_{BC} becomes zero. Physically this means that no charging current from the base terminal flows to support those required charges; in this case, the charging current associated with these flows directly from the collector to the emitter. In this case,

$$Q_{BE} = qA W_{BC} \left(R_{BE} - \frac{I_C}{qA n_{BE}} \right) = \frac{W_{BCN}}{v_E} I_C - \frac{W_{BCN}}{v_E} I_C \quad (3-55)$$

and

$$Q_{BC} = -Q_{BE} \quad (3-56)$$

The first term in (3-51), (3-53), or (3-55) is dominant at low current, and is controlled mostly by V_{BE} , for example, only through the modulation of W_{BCN} . The second term is a strong function of V_{BE} since I_C is exponentially dependent on it, and is dominant at high currents; it is slightly affected by V_{BC} also. We partition the first term to the collector terminal and the second one to the

emitter terminal (Jaffe). The buried layer (depleted) space-charge Q_{BL} at the high-low junction under the eye-collector is always supported by collector charging current.

The back-sighted hole charge Q_{BL} at the quasi-neutral emitter is a function of V_{BL} , and not influenced by the polysilicon emitter and the p-n junction. It is defined from the back-sighted hole current using a transit time parameter TE (Jaffe):

$$Q_{BL} = qA_E TE JEO \left[\exp\left(\frac{V_{BL}}{V_T}\right) - 1 \right] \quad (3-47)$$

Like Q_{BL} , the excess electron charge Q_{BLG} at the extended base is shared by the emitter and collector terminals. The energy-barrier formation at high currents affects Q_{BLG} , given by (Jaffe):

$$Q_{BLG} = \frac{q^2 N_A^2 V_T \tau_n^2}{\epsilon_0} \left[\exp\left(\frac{V_{BCO}}{V_T}\right) + \exp\left(\frac{V_{BLG}}{V_T}\right) \right] - \frac{qAN_A^2}{N_{BLG}} W_{BLG} \quad (3-48)$$

through V_{BLG} given in (3-44). The partitioning factor for this charge, f_C , is assigned a value between zero and one, analogous to (3-38).

Peripheral emitter-base junction capacitance can be significant in the highly scaled devices with small emitter area. We account for this by including a peripheral emitter-junction space-charge. To describe the peripheral emitter-junction area, we use a new model parameter WEM , which is the emitter-junction depth implied by the uniform doping profile. Assuming a cylindrical shape of side diffusion, we approximate the peripheral area as

$$A_{EP} = \pi WEM (WEM + 2R) + 4\pi WEM^2 \quad (3-49)$$

The total junction depletion charge associated with the area is then modeled by

$$Q_{JAXY} = A_{XY} \sqrt{2q\epsilon N_{AO} (V_0 - V_{AX})} \quad (3-76)$$

where V_0 is the smaller junction built-in potential.

The overall partitioned charges attributed to the emitter and collector terminals are summarized as

$$Q_{EE} = Q_E^f - Q_{AE} - Q_{AEY}^f + Q_{EX} = Q_{EEX}^f - Q_{EXXEX} \quad (3-77)$$

$$Q_{CC} = Q_C^f - Q_{CY}^f - Q_{CX} + Q_{CXX}^f - Q_{CXXCY} \quad (3-78)$$

where Q_{CXXCY} is the peripheral base-collector junction space charge [Jant95]. The substrate charge Q_{SXX} in Fig. 3-3, which accounts for the collector-substrate junction depletion, is modeled as described in [Jant95]. All the regional charge descriptions are based on the quasi-static approximation.

The charging currents in Fig. 3-4 are calculated directly without introducing capacitance. For each charge expression, the associated charging current is given by

$$\frac{dQ_i}{dt} = \frac{\partial Q_i}{\partial V_{BE}} \frac{dV_{BE}}{dt} + \frac{\partial Q_i}{\partial V_{BC}} \frac{dV_{BC}}{dt} \quad (3-79)$$

where the transconductance coefficients are evaluated from the steady state analysis. This charge-based formulation is more physical and representative of the distributed charge dynamics than is the conventional capacitance-based model such as *ChargePump*. Actual non-idealized transconductances for the BJT can be properly simulated, thereby allowing a first-order accounting for possibly significant non-quasi-static behavior [Jant95].

2.4 Summary

Significant advance in recent SiGe-base HBT technology has demanded a physical and compact device model for advanced circuit simulations. We developed in this chapter a charge-based SiGe-base HBT model, which is also applicable to an advanced silicon BJT with significant bandgap narrowing in the base region. Bandgap reduction caused by Ge incorporated into the base as well as the heavy doping in the base region are merged in a composite bandgap narrowing and exploited in a unified way. New current and charge expressions including those in [Erick5] are derived to account for the base region bandgap reduction and the related junction behavior. A comprehensive energy barrier formation model is developed in conjunction with the moving boundary conditions, which is compatible with the heterojunction device if no bandgap discontinuity exists at the emitter junction. The moving boundary conditions at both edges of the quasi-neutral base region are described in detail in terms of the bandgap variation in the base, which implicitly simulates the variation in the emitter level due to the bandgap grading. Various base current components are discussed and modeled. A network representation of this model for dc and transient simulation is given in terms of the transport currents and the regional charging currents, with proper charge partitioning. The regional charge expressions are extended to account for the bandgap variation and the energy barrier formation, and new charge partitioning schemes are discussed and suggested.

CHAPTER 4 IMPLEMENTATION OF NON-LOCAL MODEL FOR IMPACTIONIZATION CURRENT

4.1 Introduction

Accurate prediction of the breakdown voltage BV_{CEO} of a bipolar technology is important for both circuit and device designs since the system power supply is not decreased to the extent that BV_{CEO} is reduced as the technology is scaled. Conventionally the impact ionization, which defines BV_{CEO} , is modeled with a local field characterization [Sist1] of the ionization rate, $\alpha(E)$, which is based on the presumption that carriers are instantaneously compared to the steady state kinetic energy corresponding to the local electric field intensity. In a spatially-varying electric field, however, the carrier energy lags the field because of finite energy relaxation time [Sch94b]. This is the case for highly scaled BJT's in which very high electric fields and field gradients exist in the emitter-collector region. To properly model effects of impact ionization in advanced BJT's and their dependence on device structure, the ionization rate must be characterized in terms of the (average) carrier electrons for the apex BJT energy or temperature T_e (above T_c), large E , the local field (improved representation of α will overestimate the impact ionization and then underestimate BV_{CEO}).

With the objective of predictive current simulation, we develop and implement for the first time a non-local characterization for impact-ionization current in a compact but physical BJT model, applicable to both base HBT's as

well as advanced Bi JFETs. We use a previously described simplification [Shafik] of the energy-balance equation to define $T_e(x)$ from $E(x)$. This analytic model has been successfully used in numerical device simulation [Shafik] in which the electron field distribution is part of the solution. However, it has not been properly incorporated in device models for current simulation, since the conventional current models do not give a physical characterization of $E(x)$. The accounting for $T_e(x)$ in [Shafik] for bipolar current simulation is not generally valid because it uses $E(x)$ defined by the simple pre-junction depletion approximation. The field distribution in an actual bipolar transistor is more complex due to prevalent high current-induced space saturation and conductivity modulation in the collector [Jost].

The charge-based BJT model in MMSPICE [Jost, Jost, Hsieh] physically accounts for all possible charge conditions in the up collector for all bias conditions. We use it and the general $E(x)$ hypothesis as part of every transport solution for implementing the non-local model for aspect-insensitive current. This implementation is done as postprocessing in the model routine, and hence it does not significantly increase the numerical cost of the BJT model in the circuit simulator. The non-local model in MMSPICE is verified by applications to two advanced bipolar technologies, including an aggressive BiSi-base HBT process. The utility of the tool in assessing bipolar design tradeoffs involving aspect sensitivity is demonstrated by device/circuit simulations which show, for example, how variations in the BJT structure affect BF_{100} and circuit speed in scaled technologies.

4.1 Model and Implementation

4.1.1 Electron Temperature Distribution

For a given $E(x)$, $T_e(x)$ can be approximated well by the simplified energy-balance equation [256][2]:

$$\frac{d(EF_e(x) - T_G)}{dx} + \frac{T_e(x) - T_G}{l_D} = -\frac{3nE(x)}{l_D}, \quad (4-1)$$

the solution of which is

$$T_e(x) - T_G = -\frac{3n}{l_D} \int_0^x \exp\left[\frac{T_e - T}{l_D}\right] E(x) dx \quad (4-2)$$

where T_G is the lattice temperature. In this model l_D , the energy relaxation length, is assumed to be constant and, based on extensive measurements, equal to 45 nm [256][2]. To obtain particular solutions from (4-2), $E(x)$ must be modeled as well. Thus (4-2) can be directly implemented as post-processing in existing device simulators [256][2], but use of it in circuit simulations is restricted to physical device models such as the EJT model in MINIPICE [256][2, 256][3, 256][4].

4.1.2 Electron Field Distribution

Unlike empirical device models as SPICE, the charge-based EJT model in MINIPICE physically represents $E(x)$ on the *ap-catheter* for all possible conditions as part of the base-point regional analyses done to characterize the collector-transport current I_C . The collector analysis is physically linked to the base analysis, which recently we have upgraded to give a unified accounting for the electron transport in the Inp-GaAs-Inp EJT under varied β . EJT with

equivalent bandgap narrowing in the base [Hsu93]. The new heterostructural base analysis incorporates the spatially varying bandgap $E_g(x)$ as well as the doping density $N_A(x)$ directly into the electron and hole current equations, properly characterizing the effective fields acting on the minority electrons [$E_{ex} = (1/q)(dE_g/dx)$] and the majority holes [$E_{hx} = E_{ex} - (1/q)(dE_g/dx)$]. The composite model involves iterative (semi-numerical) solution of the base and collector analyses [Luo90, Hsu94].

In the (one-dimensional) collector analysis, a uniform up-doping concentration N_{D2} is assumed. Hence in any space-charge region (SCR),

$$\frac{dE}{dx} = \frac{q}{\epsilon} (N_{D2V} - n) = \frac{q}{\epsilon} \left(N_{D2V} - \frac{J_n}{q\mu_n} \right) \quad (4-18)$$

where the final approximation follows from assuming constant current in the SCR due to electron drifting at the saturated velocity v_{ns} . Based on (4-18), $E(x)$ is approximately linear, even when the depletion charge is compensated by the mobile electron charge. Depending on the base conditions, two different types of SCRs can be formed in the up-collector: a p-n junction SCR and a carrier-induced SCR associated with non-steady-state saturation [Luo87, Hsu91]. These SCRs are modulated by the position bias V_{D2} as well as J_n . In fact the entire up-collector can be space-charge, or quasi-neutral without any SCR formation as in steady-state saturation. Each case is illustrated in Fig. 4-1. Modeling of all these operational modes is important not only for the accurate description of impact-ionization current, but also for numerical stability of the EUT model [Hsu91]. The $E(x)$ distributions in terms of V_D and V_{D2} for each mode are described below. Each of them is linear, which enables analytic derivation of $T_p(x)$ from (4-12) (see also Appendix B).



Fig. 4-1 Schematic electric field patterns in the gap-collector area: (a) classical forward-satire, (b) shows quasi-saturation, (c) non-chance quasi-saturation, and (d) contemporary forward-satire operational modes.

III. Unbiased Forward Action Mode (Fig. 4.11a). In this mode, the junction SCR exists. Its width W_{SCR} can be modulated by the current as described by (4-1). The remaining neutral region of the *epi-collector* supports a constant electric field $E_{UNBIASED} = I_C/qA_pN_{ACT}$. Although this field is relatively small, it must be accounted for since T_e tends to lag E_x and hence significant impact ionization can occur in the neutral region. The electric field pattern is hence described as [Hosli]

$$E(x) = \frac{q}{2} \left(E_{ACT} - \frac{I_0}{qA_pN_{ACT}} \right) (x - W_{SCR}) - E_0 \quad 0 < x < W_{SCR} \\ = E_{UNBIASED} \quad W_{SCR} < x < W_{ACT} \quad (4-4)$$

where $x = 0$ is the collector-base metallurgical junction. The field intensity in the SCR defined by (4-4) is everywhere greater than $E_0 (= v_{cr}/\mu_n)$, below which the electron drift velocity is not saturated. Since the integral of $E(x)$ equals V_{CB} plus the equilibrium junction barrier potential ϕ_{eq} ,

$$W_{SCR} = \left[\frac{2qA_pN_{ACT} (V_{CB} - V_{BC} - \phi_{eq} E_{ACT})}{I_0 - I_0} \right]^{1/2} \quad (4-5)$$

where $E_{ACT} = W_{ACT}/qA_pN_{ACT}$ and $I_0 = qA_pN_{ACT}$.

II. Unbiased Reverse Saturation Mode (Fig. 4.11b). In this mode, the whole *epi-collector* is neutral although part of it may support a constant electric field equal to $E_{UNBIASED}$. There is no SCR formation since in this case, in general, breakdown does not occur because of the relatively low field intensity. However, in order to maintain the continuity of terminal currents across the model boundaries which define different operational modes repre-

caused by the different charge distributions in the *np*-collector (Sheet1), the electric field-distribution is needed

$$\begin{aligned} E(x) &= 0 & 0 < x < W_{\text{BET}} - W_{\text{BCHS}} \\ &= E_{\text{BCHS}} & W_{\text{BET}} - W_{\text{BCHS}} < x < W_{\text{BET}} \end{aligned} \quad (4-6)$$

The above region with W_{BCHS} is given by $(V_C - V_{CB})/E_{\text{BCHS}}$ (El-Hay, Shawa, Spies, Schumann, Mode (Fig. 4.10)). In this mode, hole push-out occurs and excess electron charge is induced to support current greater than I_{C0} , creating the current-induced SCB. The SCB is well characterized by $(n - n_0)$, reflecting uniform space charge density. Note that the polarity of the net charge is negative due to the excess electrons. The electric field in the presence of the current-induced SCB is (Sheet7)

$$\begin{aligned} E(x) &= 0 & 0 < x < W_{\text{BET}} - W_{\text{BCH}} \\ &= \frac{q}{\epsilon} \left(N_{\text{BET}} - \frac{I_C}{qA v_{d0}} \right) x - E_0 & W_{\text{BET}} - W_{\text{BCH}} < x < W_{\text{BET}} \end{aligned} \quad (4-7)$$

where again the maximum field-strength is assumed to be E_0 , consistent with the assumption of saturated carrier drift velocity within the SCB. The SCB with W_{BCH} in this case can be obtained by solving Poisson's equation (Sheet7)

$$W_{\text{BCH}} = \frac{2\epsilon v_{d0} (V_C - V_{BC})}{iE_0 v_{d0} + \sqrt{(iE_0 v_{d0})^2 + \frac{2\epsilon v_{d0}}{A} (V_C - V_{CB}) (V_C - V_{BC})}} \quad (4-8)$$

4) Conventional Forward Active Mode (Fig. 4.10d). In this (common) case, the entire *np*-collector is space-charged, and $E(x)$ is everywhere greater than E_0

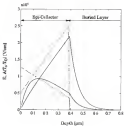


Fig. 4-2. Normalized electron temperature (solid) and electric field (dashed) distributions in the collector region, under top temporary forward-bias-free condition. The normalized electron temperature is defined by $A[T_e(x) - T_{cl}]$ with $A = k_B/k_B T_{cl}$, based on (5-4).

The space charge is positive or negative depending on α ($= I_C/\eta\phi_0\epsilon_0$) relative to N_{DPT} in (4-11), the solution of which yields (Hartl)

$$\begin{aligned} E(x) &= \frac{\eta}{\epsilon} \left(N_{DPT} - \frac{I_C}{\eta\phi_0\epsilon_0} \right) x - E_0 \quad I_C > I_0 \\ &= \frac{\eta}{\epsilon} \left(N_{DPT} - \frac{I_C}{\eta\phi_0\epsilon_0} \right) (x - W_{DPT}) - E_0 \quad I_C < I_0 \end{aligned} \quad (4-12)$$

To roughly how $T_p(x)$ looks $E(x)$, we show in Fig. 4-2 representative variations corresponding to contemporary forward-active modes of operation. As discussed in the Appendix B, $T_p(x)$ was derived analytically from (4-2) using (4-12) and $E(x) = 0$ at the biased input ($x = W_{DPT}$) in this case, and in all cases, T_p can be finite at the biased input; $T_p(x > W_{DPT})$ is derived from (4-2) with $E = 0$, using the boundary condition $T_p(W_{DPT})$ taken from the spf collector analysis. The predicted non-linear nature of $T_p(x)$ is shown in Fig. 4-3.

4.2.3 Multiplication Factor

The input-current rate α (a function of $T_p(x)$) can be derived from the experiment-based Gummel model, which is applicable for "steady state", i.e., constant T_p , conditions:

$$\alpha(E) = \alpha_0 \exp \left[-\frac{b_0}{|E|} \right] \quad (4-13)$$

with α_0 and b_0 being the proper input-injection coefficients [Skv32b]. To transform (4-13) to non-local $\alpha(T_p)$, needed for nonuniform $T_p(x)$ as implied in Fig. 4-2, a quasi-steady-state assumption is made: the steady-state ($dT_p/dx = 0$) solution of (4-14)

$$T_s(x) - T_b = \frac{k_B q_s}{k_B} (E) \quad (4-10)$$

is inserted in (4-8), and arbitrary $T_s(x)$ is allowed (Elliott). The resulting $\alpha(T_s(x))$ can now be used to characterize the nonlocal impact-ionization current. For weak generation in the *epi-collector* and the *buried layer*, the multiplication factor is

$$M - 1 = \int_{0W} \alpha(T_s(x)) dx + \int_{L_C} \alpha(T_s(x)) dx, \quad (4-11)$$

which for arbitrary RCB direction is evaluated via numerical integration using (4-8) and (4-11). The impact-ionization current is then

$$I_{bi} = (M - 1) I_C \quad (4-12)$$

4.2.3 Implementation

The implementation of the non-local model for I_{bi} is done as post-processing within the **1-TRANS** model routine (**Trans1**) of **MMSPICE**. For each bias point (V_{BE} , V_{BC}) passed in from the current model analysis, the device currents and charges are first determined without any accounting for the (weak) impact ionization. The underlying regional analysis also yield $E(x)$, from which, based on the discussion in the previous subsection and on the Appendix B, $T_s(x)$ in (4-11) and subsequently I_{bi} in (4-12) are evaluated. The computationally efficient **Runberg** integration algorithm is used for the solution of (4-11) in (4-12). The implementation in the model network is achieved by algebraically adding I_{bi} to the base current:

$$I_B \leftarrow I_B + I_{bi} \quad (4-13)$$

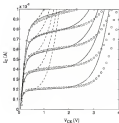


Fig. 4-3 Simulated pnpnpn-output current-voltage characteristics of NPN device with best-fit (dashed) and one-pool (solid) models for super-excitation current. Measured data are shown as circles. Base current is varied from 4 μ A to 32 μ A with 4 μ A step.

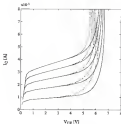


Fig. 4-8 Simulated common emitter output current-voltage characteristic of HJ device with local field (solid) and non-local (dashed) models for input-emission currents. Emission current is varied from 50 μ A to 50 μ A with 10 μ A step

4.3. Verification and Simulations

The accuracy of the non-local I_{eff} model in predicting device characteristics is assessed by application to two advanced technologies: an aggressive highly scaled SiGe-base HBT technology developed at NPT (Uguzli) (hereafter referred to as the NPT device) and a lower scaled MOSAIC-3.5 HBT technology developed at Motorola (Zohar) (hereafter referred to as the M3 device). Most of the physical device model parameters are evaluated directly from structural data such as the substrate-doping profile, the Ge mole fraction profile, and the lateral device dimensions. The substrate-region doping profile of the NPT device has a peak base doping concentration of $5 \times 10^{19} \text{ cm}^{-3}$ and a metallurgical base-region width less than 70 nm. The Ge is linearly graded across the base with 0% at the emitter junction and 8% at the collector junction. The average open-collector doping concentration is about $3 \times 10^{17} \text{ cm}^{-3}$, and the thickness $W_{\text{BCT}} = 0.4 \text{ }\mu\text{m}$. The effective emitter area used in our simulations is $2.4 \text{ }\mu\text{m}^2$. The lower scaled M3 device has a peak base doping concentration of $1 \times 10^{19} \text{ cm}^{-3}$, a base width of 160 nm, a collector doping concentration of $3 \times 10^{17} \text{ cm}^{-3}$, and $W_{\text{BCT}} = 0.6 \text{ }\mu\text{m}$. The effective emitter area of the simulated device is $1.4 \times 1.3 \text{ }\mu\text{m}^2$. To stress the accuracy of the non-local model, we also test the local-field model for I_{eff} based directly on (4-10) (which was previously implemented in MMSPICE). All the internal transport analyses are done identically for both models; only the calculations of the multiplication factors differ. For all samples/technologies, the impact-ionization parameters in (4-33) are $\alpha_{\text{ij}} = 7.03 \times 10^7 \text{ cm}^{-1}$ and $\beta_{\text{ij}} = 1.13 \times 10^8 \text{ V/cm}$, the bulk-ohm values noted as [18] [9].

The output current-voltage characteristics of both devices predicted with the local field and the non-local models, without any regional fitting, are illustrated in Figs. 4-3 and 4-4. As can be seen in Fig. 4-3, the non-local model simulates the characteristics of the NPT device well, predicting the measured

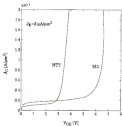


Fig. 4-3 Collector current density versus collector-emitter bias of the NVT and M2 devices calculated at the same base current density. The breakdown voltages (V_{CE0}) are taken at $J_C = 2 \text{ mA/cm}^2$.

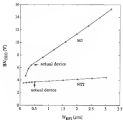


Fig. 4-6. Predicted breakdown voltages versus specific impulse for HTT and M3 technologies. The actual device characteristics [Lignoff, 82a-87] are indicated.

$BV_{CEO} \approx 1.9$ V with only about 10% error. The local-field model, however, shows significant error, predicting only about 1.5 V. The characteristics in Fig. 4-4 of the lower scaled M3 device simulated by the new local model are also in good accord with the measured data. $BV_{CEO} \approx 4.1$ V is predicted well. The local-field model prediction is again low, although the error is not as large as it is for the highly scaled device. We note that increasing λ_D from 60 nm (Ref. 14) to 80 nm, which has been proposed as representative of the underlying energy-transport physics (Chak98), results in even better accuracy of the non-local model applied in both technologies. These applications thus support our MOSFICE implementation and furthermore show clearly that non-local modeling of impact ionization is needed for highly scaled bipolar technologies.

To completely test the utility of the new local model implemented in a circuit simulator, we use MOSFICE to investigate the design tradeoff between the I_{BT} breakdown voltage and circuit speed. We vary the α -collector thickness W_{AC} in both technologies, from 0.2 μ m to 3.2 μ m in the M3 technology and from 0.1 μ m to 3.1 μ m in the NTT technology. For this design study, we estimate BV_{CEO} as the bias voltage at which the collector-current density $J_C = 2$ mA/ μ m² when the driving base current density $J_B = 1$ mA/ μ m². The representativeness of this estimation is supported in Fig. 4-5.

The predicted BV_{CEO} dependence on W_{AC} are shown in Fig. 4-6; the NTT technology shows weak dependence, while the M3 technology shows a strong correlation, with BV_{CEO} increasing with W_{AC} . These different trends in the breakdown characteristics are important in the design optimization, as we show later with reference to speed predictions. To understand the reasons for this difference we checked the operational modes where the breakdown occurs, and found that the breakdown of the NTT device happens in the presence of the junction BKN (Fig. 4-1(a)) and that of the M3 device happens in the presence of

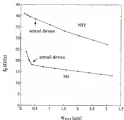


Fig. 4-7 Simulated peak cutoff frequencies versus gate oxide thickness for NIT and M3 technologies, $V_{DD} = 5.0$ V. The actual device characteristics (Lip85, Sch87) are indicated.

the current-induced SCE (Fig. 4-16c). In the BiD device then, the impact occurs first that triggers the breakdown occurs after the current gain β has been degraded due to base pushout. Since the pushout is limited by the buried layer, this degradation ceases as W_{DPT} increases. The common-emitter breakdown condition, $\beta(M-1) = 1$, then requires an increasing $(M-1)$, and hence delays an increasing BF_{CEO} with W_{DPT} . Contrarily, the breakdown-triggering impact situation in the NTF device occurs at relatively low current, and hence W_{DPT} has only a small influence on the electric field distribution since the electric drop across the buried ep-collector is negligible. Recall that the ep-collector doping level of the NTF technology is 10-times higher than that of the BiD technology. These results clearly show that accounting for the impact ionization in the junction SCE only, which is commonly done (Sze92), will not model the breakdown adequately, and that the impact situation in the current-induced SCE needs to be accounted for. Further, the insight obtained can aid optimal device design, the breakdown voltage can be controlled by varying the ep-collector thickness only when it is defined in the non-stress-squeezed extension mode, i.e., at the presence of base pushout.

To gain insight regarding base speed correlation with BF_{CEO} , we show in Fig. 4-17 MMSPICE-predicted peak cutoff frequency f_T (at $V_{CE} = 2.5$ V) versus W_{DPT} for both technologies. For the NTF technology, f_T increases significantly with decreasing W_{DPT} , which implies that the collector transit time is equally fast relative to the base and emitter transit times. This result is reasonable because the NTF emitter and base transit times are equivalently reduced due to the shallow emitter junction and the Ge grading which aids the electron drift in the base region (Sze92). For the BiD technology however there is less change in f_T with W_{DPT} , except for very thin ep, indicating predominant emitter and base transit times. The rapid increase of f_T for W_{DPT} becoming very thin is due

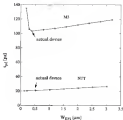


Fig. 4-8 Simulated ECL average propagation delays versus signal layer thickness for NTT and M3 technologies. The actual device characteristics (Ugoh8, Sde87) are indicated.

to the substantial reduction of the base transit time that results when the ap-collector becomes completely open-charged (Fig. 4-10d). In this mode, the quasi-neutral base-region width is modulated strongly since the proximity of the buried layer affects the depletion-region arrangement in the base. Using this insight and the fact that SV_{BESD} is a strong function of W_{BEP} in the MIT technology, we might conclude that the breakdown voltage could be increased by widening the ap-collector, without sacrificing much speed, or that the speed could be substantially increased by thinning W_{BEP} significantly, with some sacrifice in SV_{BESD} . However, design tradeoffs involving speed must be based on device architecture, and not merely on device speed, which can be misleading as we now discuss.

We use MINIPICE and the physical 3-D model to simulate the four-transistor ECL gate and to predict the dependence of propagation delay on W_{BEP} . These circuit speed simulations, correlated physically with the previously discussed device breakdown considerations, clearly show the utility of the non-local I_{BQ} model implemented in a circuit simulator. The demonstrated simulation capability afforded is essential in scaled technology design.

Fig. 4-4 shows predicted minimum ECL gate propagation delay t_{pd} versus W_{BEP} for both technologies. The MIT technology shows a slight variation in t_{pd} , slowly decreasing with decreasing W_{BEP} . However, the MIT technology shows two different trends for W_{BEP} greater than 0.4 μm ; there is a slowly but weak increase in t_{pd} with increasing W_{BEP} , in contrast to the strong $SV_{\text{BESD}}(W_{\text{BEP}})$ dependence in Fig. 4-6, but for thinner W_{BEP} , t_{pd} increases with decreasing W_{BEP} , in contrast to the speed implication of the $f_T(W_{\text{BEP}})$ dependence in Fig. 4-7. This rapid increase in t_{pd} is due to the increase of the collector junction capacitance, which is determined by the thickness of the ap-collector separating the highly doped base and buried layer. The

performance affects f_{gate} but not f_T . The insight afforded by these simulations explains why RC propagation delay does not correlate well with f_T . In fact, it correlates better with the maximum oscillation frequency of the BJT, $f_{\text{MAX}} \propto f_T^2 / (\text{FWHM}_{\text{RC}})$ [Houtzi]. Nevertheless, to effect optimal scaled bipolar technology design trading off speed and breakdown, a tool with an MINISICE-like circuit to g , i_{sat} /simulation and correlated device-to g , ΔV_{BBD} -simulation with non-local impact ionization is needed. In fact, the user here has indicated that the actual MJ device ($W_{\text{BET}} = 0.4 \mu\text{m}$) optimally effects this tradeoff with regard to W_{BET} , but that the tradeoff in the highly scaled BJT technology is not optimally sensitive to W_{BET} .

4.4 Summary

A non-local model for asymptotic carrier loss has been implemented in a physical BJT model for predictive circuit simulation. Physical characterization in the model of the electric field distribution in the open-collector for all bias points enables derivation of analytic expressions for the electron carrier temperature T_e from a simplified energy-balance equation. The carrier-multiplication factor is then calculated by numerically integrating the T_e -dependent impact-ionization rate across the collector region, including the buried layer. This non-local model is directly implemented as post-processing in the BJT model routine in MINISICE [Jassby, Jassby, Houtzi, Houtzi].

The non-local model was verified by application to two scaled bipolar technologies in which the breakdown voltages (ΔV_{BBD}), which are severely underestimated by the local field model, are predicted. The utility of the model implemented in a circuit simulator was strengthened by investigating the design tradeoff between the BJT breakdown and circuit speed as influenced by variations of the open-collector thickness (W_{BET}) in the two technologies. From the

SPICE simulations, we found that the breakdown is significantly sensitive to R_{SD} only when the triggering current multiplication comes in the current-induced ECL (in the presence of base pinches). Predictions of peak cutoff frequency (f_p) showed different dependences on W_{dECL} , depending on the degree of device scaling. However, simulations of ECL current and predictions of minimum gate propagation delay ($t_{p,d}$) revealed that f_p is not the proper basis for the design tradeoff involving the circuit speed, depending on the device structure, there can be little correlation between the device speed (f_p) and the circuit speed ($t_{p,d}$). The physical device-level simulations thereby revealed the importance of predictive circuit simulation, as well as device simulation with non-local accounting for the space-chargeless current, in topolar technology design optimization, based for example on trading off device breakdowns and circuit speed.

CHAPTER 1 HBT MODEL IMPLEMENTATION AND VERIFICATION

1.1 Introduction

MMSPICE has evolved as new features were added. MMSPICE-1.1, which is described in Chapter 2, resulted from substantial corrections to resolve convergence and computational efficiency problems in MMSPICE-1.0 [Hiro80]. It evolved further to MMSPICE-1.4 with a more physical accounting of charge partitioning of Q_{REV} , which is overviewed in Chapter 3. Then MMSPICE-2.0 was generated with the addition of new optional features for velocity overshoot, current crowding, and multi-dimensional effects [Joshi82]. MMSPICE-3.1 followed with accounting for the parasitic current through the peripheral substrate (pnp in the n-pn device structure) transistor [Chen83]. Finally MMSPICE-3.3, the present version, has evolved from MMSPICE-2.1 via the HBT modeling described in Chapters 3 and 4 (as well as the addition of a physical MOSFET model [Chen83]). The evolution of the HBTNET model has maintained complete compatibility with the previous versions.

The new features implemented into the HBT model routine of MMSPICE-3.3 are listed as follows. First, the accounting for the bandgap variation in the base region, which requires substantial modification of the device model routine for the collector transport current as described in Chapter 3. Second, the energy barrier formation model, which is implemented as a part of the transport current analysis. Third, modification of the regional charge expressions maintaining compatibility with the previous version. Fourth, a

non-local scattering for impact-ionization current based on a simplified energy-balance equation, which is implemented as post-processing as described in Chapter 4. We give a qualitative description of the implementation of all these features here.

The modeling and implementation are unified within a model parameter evaluation scheme. The linear bandgap variation and exponential doping profile assumed in our model does not exactly fit a real device, although they are representative. We develop a parameter evaluation scheme which constrains the real device structure to our model parameters, and maintains the model predictability. The evaluation scheme is based on the actual impurity doping profile and the G_0 distribution in the base. It is compatible with the homogeneous-BJT modeling, the case that there is no G_0 , but only bandgap narrowing due to heavy base doping.

We assess our model implementation by application to a highly scaled aggressive NTT Sakurbae HBT technology [179,180]. The (unshifted) intrinsic region doping profile of this technology is used to evaluate model parameters without any optimization. Results show favorable comparisons of measured and model simulated dc and ac characteristics, with and without the G_0 interpretation, which verifies that our model reliably accounts for the bandgap narrowing effect due to the G_0 as well as the heavy doping.

Through the measured and simulated ac characteristics, we found that the peak current where the maximum device speed is observed is significantly lower than the Conventional model for the base pushout. Degrading effects on EBT speed performance due to current-induced perturbation of the collector-base junction space-charge region (SCR) prior to base pushout are analyzed and assessed. Inverse base-width modulation (IBWM) – a widening of the quasi-neutral base width – and collector SCR-width widening (SCRW) as

highly scaled BJT and HBT are identified as important mechanisms governing device speed degradation at high currents.

This chapter is organized as follows. First in section 5.1, the model implementation in MMSPICE-3-D is represented qualitatively. In section 5.2, the model parameter evaluation method for the bandgap-related parameters is described. In section 5.3, representative dc and ac simulation results are compared with measured data from the mentioned technology to assess our model implementation. In section 5.4, the HBT effect is studied via MMSPICE-3-D. Finally, in section 5.5 this chapter is concluded with a summary.

5.1 Model Implementation

The new HBT analysis requires four additional model parameters, $DEGE$ and $DEEC$, the bandgap reductions at the emitter and collector metal-semiconductor junctions, respectively; $DBEL$, the high-injection minority-carrier diffusivity in the base region, and WEM , the emitter junction depth. To bring into the new model parameters through the MMSPICE input file, eight related subroutines are modified: `ADDELT`, `FIND`, `FEADON`, `SPICE`, `ALYON`, `MOODEK`, `TMPLUP`, and `QBCT` (see Programmer's Guide for MMSPICE-3-D).

To implement the mentioned new features, the previous charge-based BJT model routine, `QBCT`, is largely rewritten to reflect the new HBT model routine. A simplified flow chart of the new model routine is shown in Fig. 5-1. The new base-region transport equations are implemented in the main iteration routine, and the regional charges and base current are calculated as by-products. The energy barrier formation is integrated as a part of the main iteration routine, and the new fixed accounting of the emitter recombination current

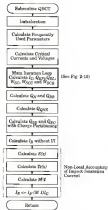


Fig. 3-1. New QMCT structure for MEMSICE-3.1

is implemented as post-processing in QRCIT. We describe here the QRCIT routine qualitatively focusing on the main three new features. One new base region carrier transport modeling, the energy barrier formation formation, and the non-local accounting of the injected-emission current.

3.2.1 Base Region Carrier Transport

The overall structure of model routine is same as in the previous version. The first part is initialization to set model parameters and various constants (including bandgap-related variables), which are used in the main iteration routine and the subsequent post-processing. Critical currents and voltages are calculated in this part. Also, to reduce the computational time in the quasi-stationary mode according to the new iteration scheme described in Chapter 2 (Section 2.3-4), some critical variables are calculated *a priori* under the assumption that the collector SCE is collapsed. Through the first iteration loop, the operating mode is identified by using the critical current `CRCCUR1` or `CRCCUR2` in QRCIT, as the first guess of the collector current. Once the quasi-stationary mode is identified, the base region analysis is done first throughout the subsequent iterations to calculate the quasi-Fermi level (QFL) separation at the collector junction, which connects the base analysis with the collector region analysis. Otherwise, the collector region analysis is done first throughout the subsequent iterations to calculate the moving boundary conditions at the collector side of the QNS region for the base region analysis. The structure of the main iteration routine is identical to that flow-charted in Fig. 3-13. The moving boundary expressions are modified substantially according to the bandgap variations in the base. The second-order charge expression in the QNS region has been changed according to the expression in (3-27), without

preserving the compatibility with the previous version. The previous version used the integral-charge-control relation for this accounting [2005a].

3.3.3 Barrier Barrier Formation

This accounting for the barrier formation applies only to the quasi-neutral mode. The critical current variable $QPSAK$ in QBCN, which corresponds to the full barrier formation in the conduction band, is calculated at each time point before starting the next iteration. Once the quasi-neutral mode is identified, the QPL separation at the collector junction is calculated first according to the new iteration algorithm described in section 2.3.4. If the QPL separation is greater than that at the other edge of the quasi-neutral region in the specification, which is represented as variable $VSCH$ set to $PC+DSCH$, base pushout is presumed. Otherwise, barrier formation is presumed, and is represented by setting the space-charge-region width to the epi-collector thickness. Increasing current levels are modeled in terms of the increased junction QPL separation with a fully depleted epi-collector region, which effectively decreases the height of potential barrier.

In accord with the implementation, the expression for the quasi-neutral region excess carrier charge $QQNR$ in QBCN has been changed. In cases of base pushout, the expression has the same form as in the previous version except for the change in the value of $VSCH$ ($-PC+DSCH$), which was set to PC . In the barrier formation mode, $QQNR$ equals zero since no quasi-neutral region exists; the modification made for this case is the separation of that operational mode from the other ones. This modification is illustrated by the $I_C - V_{BE}$ plots as shown in Fig. 3.3, which can be compared with that in Fig. 3.4 to notice the difference. The potential barrier formation also affects the electron

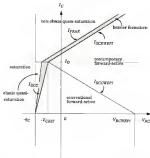


Fig. 3.3 Operational model formation in the $j - V_G$ plane for the 1D6T model in QUEST of MMS/TC-3.3.5. Potential barrier formation mode is indicated as a shaded region. See also section 3.3.

field distribution in the open-collector, which is described next in conjunction with the computing of the impact-ionization current.

3.3.1 Nonlinear Accounting of Impact-Ionization Current

The new accounting for impact ionization is implemented as post processing after the collector-current transport analysis and the regional charge calculations, but before the terminal base current is calculated as indicated in Fig. 5-1. This structure is motivated by our weak impact-ionization assumption. The first part of this routine calculates the electric field. In accord with the implementation of the potential barrier formation, we separate that mode as evaluating the field distribution. We consistently use the maximum field intensity of $E_0 (= v_{ps}/\mu_{p,eff})$ within the open charge region as described in Chapters 3 and 5. The collector region is presumed to comprise three separate regions: open-charge region, quasi-neutral region, and the buried layer, where the first two are unified to the open-collector region. The electric field in the buried layer is assumed to be negligible in the present model implementation. Five variables are set accordingly (see Appendix B) to describe the field distribution for the subsequent electron temperature calculation.

The electron temperature is then calculated analytically for each subdivided collector region in terms of the five variables. Each expression is integrated numerically for each region using Romberg's algorithm and is summed to yield the total impact-ionization multiplication factor $M-1$, as described in Chapter 4 and Appendix B. For the terminal base current calculation, the impact-ionization current, $I_B(M-1)$, is subtracted from the base current calculated without considering the impact-ionization to give the net terminal current. This formation is valid only for weak impact ionization, which is the usual mode of interest, for example to predict BF_{CEG} .

3.2 Bandgap Parameter Evaluation

3.2.1 DEGR and DRGC Evaluation

In our model, the bandgap profile in the base region is represented as a linear function represented with two new model parameters $DEGR$ and $DRGC$. $DEGR$ is the model equivalent of the total bandgap reduction at the emitter metallurgical junction and $DRGC$ is that at the collector metallurgical junction. Note that our model assumes an exponential doping profile, while the actual doping profile is not necessarily exponential. In scaled BJT devices, the exponential representation of the base profile is not accurate. In representing the actual doping profile with an exponential profile, the equivalence of both profiles with respect to the base-Gummel number and the averaged built-in field along on the carriers should be maintained. In BJT devices for a highly scaled silicon device with significant bandgap narrowing in the base; however, the emitter band edge and the valence band edge are not necessarily parallel with each other, which implies a difference between the built-in collector/fields along on the minority and majority carriers. The effective base Gummel number, which describes the base region transport current [Kryzll] and underlies the minority-carrier transport, depends on the bandgap variation in the base region and is different from the (unmoderated) base Gummel number. Thus this effective base-Gummel number and the effective built-in field it defines need to be matched in representing the effective doping profile in the model.

The first step in evaluating $DEGR$ and $DRGC$ is to transform the actual doping profile into an equivalent exponential profile. Based on negligible majority-hole current in an n -pn device, the built-in field E_p along on the majority holes can be approximated as (see Appendix A)

$$E_p = \frac{-kT}{q(N_{D0} - n_{D0})} [\ln(N_A(x_{D0})) - \ln(N_A(x_{DD}))] \quad (5-1)$$

where the two QSD boundary edges x_{D0} and x_{DD} can be obtained by solving the one-dimensional Poisson's equation across the junction GCIs under the thermal equilibrium condition. Note that the calculation of n_{D0} and x_{D0} also reflects the bandgap variation in the GCIs. Since the band is field obtained from the equivalent model exponential profile, $N_A(x) = N_{A0}\exp(-qx/W_{\text{SGT}})$, is

$$E_p = \frac{qkT}{qN_{A0}}, \quad (5-2)$$

the factor q is evaluated by equating (5-1) and (5-2). Then equating the integration of the actual profile with that of the exponential profile across the base region yields N_{A0} .

The next step is to obtain an effective doping profile $N_A'(x)$ from the actual doping profile according to the following relation:

$$N_A'(x) = N_A(x) \exp\left(-\frac{\Delta E_g(x)}{kT}\right) \quad (5-3)$$

where $\Delta E_g(x)$ is the total bandgap reduction at x . Then, since $\Delta E_g(x)$ is assumed to be linear, another model exponential profile for the effective doping profile, which can be represented with N_{A0}' and η' , is defined. The effective built-in field acting on the minority electrons is then approximated with (see Appendix A)

$$E_n = \frac{-kT}{q(N_{D0}' - n_{D0}')} [\ln(N_A'(x_{D0})) - \ln(N_A'(x_{DD}'))], \quad (5-4)$$

and the evaluation of N_{A0}' and η' is done in a way similar to that used in evaluating N_{A0} and η .

The values obtained for M_{AB} and η , which are model parameters, are used directly. However, M_{AB}' and η' must be used (with M_{AB} and η) to evaluate the model parameters $DEOE$ and $DEOC'$. Assuming that (5-3) is applicable in the QRS, we can at to define

$$DEOE = kT \ln \left(\frac{KAD}{NAG} \right) \quad (5-3)$$

and

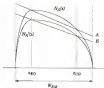
$$DEOC' = kT \ln \left[\frac{NAG \exp(-\eta)}{KAD \exp(-\eta')} \right] \quad (5-4)$$

The evaluation procedures described so far are illustrated in Fig. 5-3 schematically.

5.3.3 Model Parameters Affected by Banding Variation

The evaluation of the moving boundaries x_1 and x_2 is done by solving the one-dimensional Poisson's equation across the junction SCS. It is noted that since the SCS alloy has different material characteristics from the silicon, the junction SCS behavior and thus the moving boundaries should reflect the change in the permittivity in solving Poisson's equation. The permittivity of the SCS alloy, however, is not clearly known as a function of the Ge mole fraction. To overcome this difficulty, we assume the permittivity of SCS alloy can be linearly interpolated between those of silicon and germanium. This should be reasonable since the Ge concentration incorporated into silicon is relatively low.

Integrating Poisson's equation across the entire junction then yields the built-in potential PB , a model parameter, as follows



1. Transform $N_A(x)$ into $NADexp(-\eta x/W_{ABP})$
2. Transform $N_A^*(x)$ into $N_A^{*'}(x)$
3. Transform $N_A^{*'}(x)$ into $NACFexp(-\eta' x/W_{ABP})$
4. Evaluate DGE and DGC

$$DGE = ATrn(NAD/NACF)$$

$$DGC = ATrn(NAD/NACF) - RT(\eta - \eta')$$

Fig. 8-2. Binding parameter evaluation scheme

$$\begin{aligned}
 PE &= - \int_{x_{BE}}^{x_{BB}} E(x) dx \\
 &= -x_{BB} E(x_{BB}) - x_{BE} E(-x_{BE}) - \int_{x_{BE}}^{x_{BB}} x \frac{dE(x)}{dx} dx
 \end{aligned} \quad (3-7)$$

$$= \int_{x_{BE}}^{x_{BB}} E \frac{qN_A(x)}{x_{BB}} dx - \int_{x_{BE}}^{x_{BB}} x \frac{qN_A(x)}{x_{BB}^2} dx \quad (3-8)$$

where the sign of x is at the emitter metallurgical position, and $-x_{BE}$ and x_{BB} are the emitter- and base-side edges of the emitter base junction SCE, respectively. PE is evaluated by equating the following to (3-8)

$$PE = \frac{1}{q} (E_B(x_{BB}) - E_{BE}) + \frac{kT}{q} \ln \left(\frac{N_A(x_{BB})}{N_A(-x_{BE})} \right) \quad (3-9)$$

where $E_B(x_{BB})$ is the bandgap at the position x_{BB} . The collector junction built-in potential PC , also a model parameter, is similarly evaluated. The boundary x_{BB} , derived from Poisson's equation at thermal equilibrium, is transformed into WSE0, the model parameter used in describing the emitter-side moving boundary edge x_E of the QWB. Then, equating the total depleted charge of the emitter junction SCE in the base with reference to the actual dopant profile in that with reference to the transformed exponential profile, the model parameter AE is evaluated in conjunction with WSE0 [Dro03].

3.4. Verification of Modeling

The model in HSPICE-3.0 has been verified based on an aggressive SiGe-base HBT technology, which uses MBE to grow the alloy-base layer and selective W CVD to form the base electrodes [Ug96]. The baseline device has a base width of less than 70 nm with a peak boron doping concentration of about $5 \times 10^{18} \text{ cm}^{-3}$ as revealed by the plot of the intrinsic doping profile in Fig. 3-4. The Ge is linearly graded up from 0% at the emitter junction to 8% at the

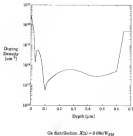


Fig. 3-4 Intrusion doping density profile of the baseline transistor of the GaGe beam LED technology

collector junction. The device structure was used to evaluate the model parameters, with minimal tuning for the physical parameters related to the quantum interface states between the polysilicon and bulk emitter regions. Test devices with HBT and without (Si BJT) the Ge intergrowth were constructed and simulated.

Figs. 5-5 to 5-6 show the device-level comparisons of measured and model-simulated device characteristics for both the HBT and Si BJT. The Gummel plots in Fig. 5-5 represent devices with $0.2\text{e}3\text{ }\mu\text{m}^2$ emitter areas; the $f_T(I_C)$ and $f_{MAX}(I_C)$ characteristics in Figs. 5-7 and 5-8 represent two such devices in parallel. The $I_C(V_{BE})$ comparisons (Fig. 5-5) indicate that our scaled base model assumes well for the Si bandgap narrowing and as well as for the SiGe heterostructure. The HBT peaks of $f_T(I_C)$ (Fig. 5-6) and $f_{MAX}(I_C)$ (Fig. 5-7) at about 17 GHz are predicted well as are those of the Si BJT. Our simulations reveal that the peripheral emitter-base junction capacitance and the intrinsic emitter-base junction capacitance in the scaled HBT significantly limit f_T and f_{MAX} at low and intermediate I_C . For example, the predicted peak value of f_T without these parasitic capacitances is about 30% higher.

Our simulations further suggest that the HBT performance is very sensitive to the polysilicon-emitter contact structure, which implies that the emitter transit time is comparable to the base transit time. Fig. 5-8 shows how $f_T(I_C)$ is improved as the Ge mole fraction at the collector junction is increased to 30%. These predictions imply that the base transit time can be reduced, but that the speed performance is ultimately limited by the emitter delay, due to charge stored in the polysilicon contact as well as the heavily doped Si. Simulations of f_{MAX} give the same insight.

HBT ECL-circuit simulations, which are obtained from simulations of a four transistor gate depicted in Fig. 5-9, faithfully predict gate propagation

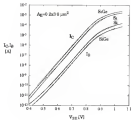


Fig. 5-6: Measured (dashed) and simulated (solid) Gummel plots, with and without the incorporated. The emitter area is $0.2 \times 0.2 \mu\text{m}^2$.

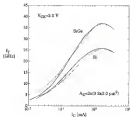


Fig. 8-8 Measured (dashed) and simulated (solid) output frequencies, with and without the incorporated. The DFT is two transistors in parallel, each with emitter area of $0.2 \times 0.6 \mu\text{m}^2$.

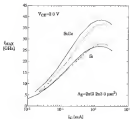


Fig. 8.7 Maximum (dashed) and simulated (solid) power-gain and max. frequency, with and without the compensation. The DUT is two transistors in parallel, each with carrier area of $0.2 \times 0.2 \mu\text{m}^2$.

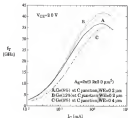


Fig. 3-8 Simulated cutoff frequencies for increased Ga concentration at the junction and for a wider emitter stripe. The base-emitter results (solid) are included for comparison.

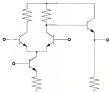


Fig. 5-9 One stage of four transistor standard ECL gate circuit.

delay, which is reduced to less than 20-ps with increasing switching current as shown in Fig. 5-12. Corresponding results for the 3n-THT ECL circuit are also shown. To illustrate the predictability of our model, and to give physical insight regarding the speed performance of the SiGe base HBT technology, we doubled the 1-4 μm emitter width (and area) and increased to 120% the Ge mole fraction at the collector junction, and re-simulated the ECL gate. The results in Fig. 5-13 show how Geas device changes affect the gate propagation delay. The predicted decrease in delay yielded by the higher Ge concentration is minimal in accord with the discussion of Fig. 5-4. The wider emitter seems to result in a relative loss of speed, which the simulations suggest is a result of the larger intrinsic base resistance R_{bi} . Interestingly this prodigious loss of speed is clearly reflected by corresponding simulations of f_{Tmax} for the larger HBT, but its degree is not indicated by the f_T predictions, which are included in Fig. 5-4.

However, we note that although the model accounts for the bias dependence of R_{bi} , it lumps it with the intrinsic base resistance [Eq.(5a)]. This is non-physical, and can overly emphasize base constants defined by R_{bi} and the parasitic capacitances decreased previously. We therefore re-simulated the wider-emitter ECL circuit, accounting for the BE dependence of the parasitic capacitances but ignoring the BE dependence of R_{bi} . The predicted gate propagation delay from this simulation, which is included in Fig. 5-13, still shows no benefit from the increased BE. Knowing that the actual increase in R_{bi} (which is proportional to W_{BE}) would cause some speed loss, we conclude that increasing BE to reduce the parasitic capacitances has significant would in fact degrade the speed performance of the HBT ECL circuit.

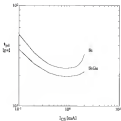


Fig. 6-35 HBT/BCJ gate propagation delay derived from simulations of four-transistor circuit, compared with delay for the Si BJT complementary circuit. The transistors have emitter areas of $0.2 \times 0.2 \mu\text{m}^2$.

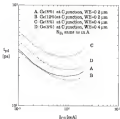


Fig. 5-11. Predicted HBT RCCL gate propagation delays for increased Ge concentrations at the collector junction and for a wider emitter stripe. The baseline device results (noted) are included for comparison.

2.5 Inverse Base-Width Modulation and Collector-Space-Charge Region Widening: Increasing Effects at High Current Densities in HBTs

Base-pushing (RCEB), or non-classic quasi-saturation, is the usual mechanism underlying the degradation of speed performance at high collector currents (I_C) in contemporary HBTs. At high-current densities, mobile carriers modulate the collector-base junction space charge, and ultimately push it into the optimal collector away from the metallurgical junction. Thus the quasi-saturated base region is effectively extended into the collector, causing increased carrier transit time and degradation of device speed, e.g., of the cutoff frequency f_T . In highly scaled devices, the collector doping concentration is increased and/or the optimal thickness is decreased to form the onset of base-pushing at higher-current densities and to minimize its effect on device speed. The pronounced design trade-off is retained: high-speed performance at higher currents, but with a lower breakdown voltage.

However, when the current density is increased in highly scaled devices, two other mechanisms associated with the modulated collector-base junction space charge can cause speed degradation. These mechanisms, which have heretofore been partially ignored, are "inverse" base-width modulation (IBWM) and collector-space-charge region width widening (RCRW), and they occur at current densities well below the onset of base-pushing. IBWM is a widening of the quasi-saturated base width that results when the collector-base junction SCR on the base side shrinks as I_C increases and mobile carriers modulate the space charge and diminish the electric field at the junction. The consequent increase in the base transit time will degrade f_T if indeed the transit time is important in limiting f_T . In highly scaled HBTs and HBTs [Cp100], MMSPICE 2.2 simulations show that this is the case [Hos03], and hence the peak f_T - I_C and general speed performance are limited by IBWM

and not base pushout. In lower scaled devices, IBSW occurs at lower current densities where delays associated with the emitter-base junction overwhelm its effect, and hence the peak f_T - f_C is limited by base pushout at somewhat higher currents. In further scaled devices, MMSPICE-3.0 simulations predict that the increase in the collector BCJ transit time due to BCJW will degrade f_T , and indeed can be the predominant mechanism degrading speed at high currents.

In this section, we describe IBSW and BCJW analytically, and we show how IBSW is responsible for f_T degradation in highly scaled BJTs from an aggressive SiC/Si base technology (Table IV). The analysis is supported by device measurements and by numerical device-level simulations using MMSPICE-3.0. We also discuss, based on MMSPICE-3.0 simulations, how BCJW can define maximum f_T in further scaled devices.

3.3.1 IBSW and BCJW

The base widening due to IBSW and BCJW are illustrated in Fig. 5-12. For I_C high enough that the electron density (in $\text{W} \approx I_C/qA v_{sat}$, where v_{sat} is the saturated drift velocity) is comparable to, yet less than the doping density N_{dsc} in the epitaxial (collector) layer, the electric field in the collector-base junction (CBJ) is perturbed as shown. The magnitude of the peak electric field, E_{pk} , decreases with increasing I_C because of the current-induced perturbation as well as the sheet drops in minimum and maximum collector resistance. This causes the width of the BCJ on the base side of the junction, W_{dsc} , to decrease and hence the width of the quasi-neutral base, W_B , to increase, leading to limit f_T and device speed performance. It is obvious that IBSW begins at values of I_C well below the onset current for base pushout, which is

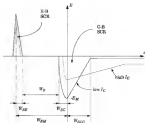


Fig. 6.12 Typical electric field profile in an edge-triplet transistor, for low and high currents, defining several widths used in the IFM and SCF analysis.

higher than the saturated drift velocity limited current, $I_D = qA\mu_{eff}N_{eff}$ [Jacobitz]

To express $\mathcal{W}_{BC}$ in terms of I_C , we first relate $\mathcal{W}_{BC}$ to E_B . Assuming an exponential base doping profile $N_B(x) = N_{eff}\exp(-\eta x/\mathcal{W}_{BC})$ as in [Jacobitz] and neglecting the electron density in the highly doped base side of the BCN, we can express $\mathcal{W}_{BC}$ via Poisson's equation,

$$\mathcal{W}_{BC} = \frac{\mathcal{W}_{BC}E_B}{\eta} \ln \left(\frac{e\eta E_B}{(2\mathcal{W}_{BC})^2 N_{eff}} + 1 \right) \quad (3-10)$$

where $N_{eff} = N_{eff}\exp(-\eta)$. Assuming that the integral of the electric field across the entire epi-collector region equals the junction built-in potential, ϕ_C , plus the reverse-biased BCN bias on the junction, $-\mathcal{V}_{BCN}$, we next express E_B in terms of I_C via Poisson's equation [Jacobitz],

$$E_B = \left[\frac{2(I_D - I_C)(\phi_C - \mathcal{V}_{BCN} - I_C R_{BCN})}{eA\mu_B} \right]^{\frac{1}{2}} + \frac{I_C R_{BCN}}{\mathcal{W}_{BCN}} \quad (3-11)$$

where $R_{BCN} = \mathcal{W}_{BCN}/(e\mu_B N_{eff})$. Note that the external reverse bias is $-\mathcal{V}_{BCN} = -\mathcal{V}_{BCN} + \phi_C R_C + I_C R_B$, where R_C and R_B are the external collector and base resistances, respectively. Related to (3-11) is the junction SCE with on the collector side, $\mathcal{W}_{BCN}$, which is expressed as [Jacobitz]

$$\mathcal{W}_{BCN} = \left[\frac{2\mu_B A_C (\phi_C - \mathcal{V}_{BCN} - I_C R_{BCN})}{I_C - I_C} \right]^{\frac{1}{2}} \quad (3-12)$$

Note that $\mathcal{W}_{BCN}$ might increase or decrease as I_C increases (depending on $\mathcal{V}_{BCN}$) [Jacobitz], and that (3-11) and (3-12) are valid only when $\phi\mathcal{W}_{BCN} > \mathcal{W}_{BCN}$.

The combination of (3-10) and (3-11) defines $\mathcal{W}_{BC}(I_C)$. In the low-current region ($I_C \rightarrow 0$), the terms including I_C in (3-11) are negligibly small,

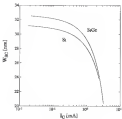
and R_{BE} hence W_{BE} are virtually constant, independent of I_C . For moderate currents ($I_C < I_D$, but comparable), R_{BE} decreases sharply with increasing I_D , and W_{BE} decreases according to (3-16). The decrease of W_{BE} weakens BSWM (see Fig. 3-1E). $W_B = W_{BE} + W_{BC} + W_{BC'}$, where W_{BC} is the width of the emitter-base junction BCB on the base side (which decreases slightly with increasing I_D , or W_{BE} , as well). Since the base transit time τ_b is proportional to W_B^2 [4 = 1 - 12] where k depends on the built-in electric field for electrons in the quasi-neutral base [Kroth], the above equations characterize how τ_b increases with increasing I_D due to BSWM. In contemporary, lower scaled devices, BSWM occurs at relatively low I_D because M_{BEY} coefficients I_D are low. At low currents, the emitter capacitance charging time is the predominant delay in the BJT (Sze61), and thus the effect of BSWM is obscured. However, in highly scaled devices, M_{BEY} is one or two orders of magnitude higher, and BSWM occurs at higher I_D where the emitter delay is significant (Sze61). Furthermore, the higher M_{BEY} means that R_{BE} and W_{BE} are larger, in accord with (3-16) and (3-11), and hence BSWM is more significant. It lengthens τ_b , and it occurs at currents for which τ_b can limit speed performance, even though W_{BE} is narrow in the highly scaled device. Indeed BSWM can weaken the high- I_C degradation of f_T in advanced BJTs.

When $-V_{BEY}$ is sufficiently large ($> I_D R_{BEY} = \phi_B$), $W_{BC'}$, as described by (3-12), increases as I_D increases. Thus in addition to BSWM, BCBSY will tend to degrade f_T prior to base pushout as well because the collector BCB transit time ($W_{BC'}/v_{sA}$) (Sze17) increases proportionally. In highly scaled devices with extremely thin W_B , BCBSY can be the predominant f_T -degrading mechanism.

5.5.2 Simulations and Measurements

We verify and demonstrate the importance of BWM in highly scaled transistors by analyzing the aggressive SiGe-base HBT technology described in sections 5.4 (Uysal2). Test devices with HBT (and without Si BiT) SiGe test periods were simulated and measured. The Si BiT is different from the SiGe HBT in three respects: it has no Ge in the base, it has a slightly different base doping profile, and it has a larger N_{dop} (e.g. $1 \times 10^{17} \text{ cm}^{-3}$). Device measurements are interpreted with the help of MMSPICE-3.1.

Fig. 5-4 shows the simulated and measured $f_T(I_C)$ characteristics of two transistors in parallel, each with an effective emitter area of $0.3 \times 3.3 \mu\text{m}^2$ at $V_{\text{CE}} = 3 \text{ V}$. The simulation profile well, for both the SiGe HBT and Si BiT, the peak f_T values and the currents at which they occur. Examination of these current values (about 2 mA) reveals that they are less than one third I_{CQ} for the SiGe HBT and less than one-sixth I_{CQ} for the Si BiT in the two-transistor test structure, meaning that subthreshold speed degradation occurs well below the currents at which base-pulsed measurements. Corroboration that BWM underlies this surprising result is given in Fig. 5-11 where W_{BWC} , obtained from the MMSPICE simulations, is plotted versus I_C for both the HBT and the Si BiT in the two-transistor test structure. Although the low- I_C values of W_{BWC} differ between the two devices because of slight differences in base doping near the collector junction(s), note that W_{BWC} begins to decrease significantly at $I_C \approx 2 \text{ mA}$ in both devices. These results are consistent with the measured (and simulated) f_T fall-off with increasing I_C shown in Fig. 5-4, and thus support the significance of BWM in these highly scaled transistors. The MMSPICE simulations further reveal only small variations of $W_{\text{BWC}}(I_C)$, implying that the BWM-based degradation of f_T is negligible here. As implied by (5-18), the



$$W_{HB} = W_{HB0} - W_{HB1} \cdot f_H$$

with high f_H degradation

Fig. 8-16 Inverse Base-Width Modulation (IBWM) illustrated for both Si and SiGe base-emitter devices

is the case because $-V_{GG}$ is not sufficiently large and R_{GG} is not sufficiently small.

To show how SCRW can limit device speed, we considered a further scaled SiGe HBT in which the Ge grading is disabled, W_{BB} is decreased by one-third, retaining the same base Gummel number, and W_{GG} is halved, which halves R_{GG} . In this device, the DFTB effect is lessened because the higher base doping density suppresses N_{DE} even at low currents, and because the narrower W_{BB} enlarges Ge grading under the base transit time comparable to the collector transit time. Hence the SCRW effect is more predominant. The MINIFETC predicted dependence on V_{GG} of maximum f_T (f_{Tmax}) and the current (I_{Cmax}) at which $f_T = f_{Tmax}$ for this scaled HBT confirmed that would not. Especially for higher voltages ($V_{GG} \geq 0$ V), the SCW transit time (W_{GG}/v_{tr}) and SCRW are predominant in determining f_{Tmax} and I_{Cmax} is determined by SCRW. Therefore, both f_{Tmax} and I_{Cmax} were predicted to decrease as V_{GG} was increased.

3.6 Summary

In this chapter, we described our HBT model implementation in MINIFETC-3.0 qualitatively, and we demonstrate systematic and self-consistent model parameter evaluation scheme considering the bandgap variation in the base region. The model implementation was demonstrated based on an aggressive 300-nm SiGe HBT and a 65-nm HBT fabricated at NIT. The MINIFETC-3.0 simulated and measured dc and ac device characteristics are in good agreement, which proves the practical utility of our HBT/HBT model. Inverse base-width modulation (IBWM) and collector space-charge region widening (CSRWO) were discussed analytically and identified as the predominant mechanisms underlying speed (f_T) degradation at high current densities in highly scaled

Bi/Te and HfTe. Measurements and simulations of the aggressive BiTe base technology showed that IBWM causes severe degradation of f_T at currents much less than the rated current for base pushout. Simulations of a further scaled HBT showed that SCRW potentially stabilizes f_T degradation in future high-speed devices. In these devices, the collector doping density is high, and hence IBWM and SCRW occur at currents high enough that the emitter capacitance charging time is negligible, and hence are despicable, unlike in lower scaled devices. This revelation means that the transistor-level design optimization of advanced, highly scaled BiT (and HBT) technologies for high-current performance must take IBWM and SCRW into account, and that basing design on base pushout alone can lead to overly optimistic, invalid speed/performance expectations.

CHAPTER 5 SiGe HBT DESIGN USING MNMSPICE-3.0

5.1 Introduction

The design flexibility provided by the Ge incorporated into the base region of the SiGe HBT requires new reliable design tools to exploit its full utility as IC TCAD. The conventional approach to mixed mode device/circuit simulation for IC TCAD, which integrates a commercial device simulator with a circuit simulator, is unduly inefficient and expensive. Alternatively, the physical SiGe-HBT model described in Chapter 3, implemented in MNMSPICE-3.0, provides a reliable and efficient simulation capability for device as well as circuit design, as demonstrated in Chapters 4 and 5.

However, we have not explored fully the utility of our HBT model in MNMSPICE-3.0 as a design-aid tool. Since, unlike the conventional homogeneous BJT, the bandgap profile in the base region can be controlled independently from the other structural parameters, it will be very instructive to explore the performance limit of the SiGe-based HBT technology implied by optimal Ge distribution.

With this objective, and wanting to demonstrate the utility of our HBT model in MNMSPICE-3.0 as a TCAD tool, we use it here to design a high performance SiGe-based HBT for digital, analog, and mixed-mode IC applications. We start the simulation exercise with reference to the actual technology developed at NTT (Fig. 32) mentioned in the previous chapters. However, we consider a different base structure with uniform doping for two main reasons. First,

bandgap narrowing due to heavy doping tends to diminish the aiding built-in electric field due to the doping gradient. Indeed our simulations of the NIT device presented previously revealed negligible benefit from the built-in field. Second, epitaxial technology, with constant doping, is advantageous for SiGe HBTs. The SiGe alloy grows via epitaxy as a layer and more compatible with conventional silicon IC technology, and the Ge distribution is easier to control. In the design exercises, we compare optimized device characteristics with published data to support the model validity as a design-aid tool for new technology development, as well as for IC design.

This chapter is organized as follows. In section 8-2, we discuss various design considerations qualitatively to gain insight. In section 8-3, we systematically effect design optimization for mixed mode circuit applications using MINITAB 8-3 for both device and circuit simulations. Simulation results are compared with a representative contemporary device published in the literature to give the TCAD exercise credence. Finally, in section 8-4, this chapter is concluded with a summary.

8.2 Design Considerations

There are fundamental restrictions in SiGe HBT design. First, the high current gain achievable in this technology is not always desirable since it tends to lower the common-emitter breakdown. Thus the allowed maximum current gain is limited by the required minimum breakdown. Second, the maximum Ge grading allowed for a given base width is limited by the stability of the strained SiGe layer [Mat94]. Third, the peak base doping density cannot be increased beyond a certain limit because of emitter-base junction forward-biasing. Finally, to achieve higher device or (digital) circuit speed, the resistivity of the emitter base region (R_{EB}) should be minimized, which virtually entails silicon films on the

base region width, gives the upper limit on the peak base doping density. In Ga HBT design, under these conditions it is challenging that a systematic approach based on a tool like MINIPACK-3.0 is called for.

To design a high performance device for mixed-mode analog applications, we want to maximize the common emitter breakdown BV_{CE0} , the Early voltage V_A , the current gain β , the unity current gain frequency f_T , and the unity power gain frequency f_{MAX} , and minimize the BCL propagation delay t_{pCL} . In digital applications only, V_A and β are not as critical, however, f_{MAX} and t_{pCL} are most important figures of merit in achieving fast current speed. In analog applications, V_A and β are more important. Since some of these performance parameters are severely correlated, design tradeoffs are inevitable.

The most fundamental restriction on the device design is imposed by the power supply voltage, which sets the lower limit of BV_{CE0} . Once this is determined, β should be traded off with the collector region parameters, the epitaxial doping concentration N_{EP} , and its thickness W_{EP} . Higher β (100-200 is preferred) is generally desirable for analog applications since it reduces the base current level. With the emitter-region recombination characteristics fixed, β is defined by the effective base Gummel number, with which it varies inversely.

The Ge grading in the base region, which is a unique flexibility in the SiGe HBT technology, can reduce the base transit time and thus enhance f_T significantly by providing an aiding field for minority-carrier transport. However, the steepness of the Ge grading selected by the strained-layer stability criterion [MacM], which dictates the maximum total amount of Ge in the layer. The steepest grading can be achieved by reducing the base region width W_{BB} , using the maximum allowed amount of Ge. However, the base width reduction increases R_{BC} , and hence degrades f_{MAX} and thus digital perfor-

means. Also increasing the Ge content tends to reduce the effective base Gummel number, which implies higher base doping needed to realize low enough β . Assuming the peak doping level is eventually restricted by the tunneling current at the emitter junction, we conclude there is a definite limit on the maximum W_{BB} , and thus on the maximum Ge grading. This consideration is important in exploring the device speed performance limit since the speed can be defined predominantly by the base-region carrier transport.

One of the unique advantages of the SiGe HBT technology is the large Early voltage achievable, which is needed for large small-signal output impedances. It can be enhanced substantially over the homojunction BJT by the graded-Ge. Under the total amount-of-Ge limitation, maximum V_A is achieved by increasing Ge at the collector junction (PvN04), which is consistent with maximum Ge grading toward the collector junction. Thus Ge grading provides two significant advantages: base transit time reduction and enhancement of Early voltage. We can surmise from this insight that the best grading scheme should be triangular since it allows maximum Ge at the collector junction subject to the total Ge content limitation (Hart90) (see Appendix C). Therefore, for the uniformly doped *npn* base HBT, the significant design parameters are base doping density, metallurgical base-region width, *npn*-collector doping density, *npn* collector thickness, and the Ge content at the collector junction, as illustrated in Fig. 4-1.

6.3 Device Optimization for Mixed-Mode Circuit Applications

We focus on the intrinsic device structure, assuming the same lateral discussion as in the NTE HBT. We assume all the emitter structures in our study are identical, in other words, the emitter formation can be done independently from the other processing steps. This assumption is realistic since low-

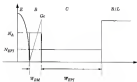


Fig. 4-1. Design parameters for optimization of BiGe_{0.5}Se_{0.5}-based HCTF

temperature and/or short-time (e.g., rapid thermal annealing) processes are commonly used for the emitter annealing/diffusion such that the pre-existing device structure is not redefined significantly. This assumption does not restrict the general conclusion of our design exercise. Accordingly, the four emitter-region model parameters (IE , $JE0$, $JE0Q$, and NEN) are not changed in our simulations. We further allow for the same degree of parasitics associated with the external device regions in the NPT device. The emitter area of the baseline device is $8.8 \mu\text{m}^2$ ($\approx 0.2 \times 0.3 \mu\text{m}^2$), and the width of base-emitter region (LEB) is $0.3 \mu\text{m}$. The (refined) emitter junction depth (REFM) is fixed at 30 nm .

We start from the NPT technology and optimally design devices for mixed-mode circuit applications. As noted in Chapter 4, the NPT device shows a significant inverse-base-width modulation (IBWIM) effect, which is caused by the low doping density at the collector-side of the narrow base region as revealed in Fig. 4-8. Designing for higher doping density at the collector-side of the base region, realized in the opt-base technology, can reduce IBWIM significantly. The higher doping at the collector-side of the base can also increase the Early voltage, which is an important figure of merit in analog applications. Low-temperature epitaxial-base formation with uniform doping is attractive in these respects due to the suppressed *x*-*y* diffusion of base dopants into the epitaxial collector during the base-layer growth.

The requirements of the device optimization for high performance mixed-mode circuits are: (i) the Early voltage needs to be improved (initially Calsonic 480N) to obtain a high intrinsic voltage gain, which is important to reduce the loading effect at the collector terminal, (ii) the current gain needs to be increased (at least above 100), but the breakdown voltage should be kept at least at the same value ($\approx 3.5 \text{ V}$). With these improvements, the ac character-

which should not be degraded relative to those of the NPT device. With the objective of optimally trading off the figure-of-merit, we simulate various device structures using MINIPICE-3.0. The NPT model parameters are evaluated from projected doping profiles as obtained in Chapter 3. These evaluations, analogous to the methodologies used in SLIMM [Krolick] for the NPT model as earlier versions of MINIPICE, ensure proper correlations among the model parameters. The doping profiles assume the same emitter structure, and uniform doping in the epi base and epi-collector regions.

The first column in Table 4-1 reflects the MINIPICE-3.0 simulation of the actual NPT technology (see Chapters 4 and 5). It shows a very low V_A , which makes it unsuitable for analog applications. We note that V_A in our simulations extrapolated at the bias point where $I_B = 1 \mu\text{A}$ and $V_{BE} = 0 \text{ V}$. The second column reflects an optimistic version of the NPT device, defined to have exactly the same β and base Gummel number B (hence higher V_A because of the higher base doping near the collector junction). The last two columns in Table 4-1 reflect attempts to increase V_A further by increasing N_A , but keeping N_D fixed to maintain β and B . Hence W_B is varied inversely with N_A . Note that the inherent bandgap narrowing due to heavy doping in the base leads to less Ge in the base to maintain the same β for increasing N_A . This required adjustment in fact results in V_A actually decreasing with increasing N_A . Note that $I_{B,CE}$ and I_{CE} show little variation, although there is a noticeable but non-monotonic change in I_F . The breakdown voltage is almost constant as expected with little variation in β . From these results in Table 4-1, we conclude that the constraint of constant N_D prohibits achieving the desired V_A . The primary reason for this is that the Ge content at the collector junction must be lowered, which tends to decrease V_A (see Appendix C for explanation).

Table 6-1 Device Characteristics for Nearly Constant Current Gain and Base Current Number

Device(Ge)	0.1%	0.7%	0.9%	0-4.2%
$W_{\text{base}}(\mu\text{m})$	63	48	35	31
$N_{\text{A}}(\text{cm}^{-3})$	(N/T)	5×10^{18}	4×10^{18}	5×10^{18}
β	49	30	34	61
$W_{\text{emitter}}(\mu\text{m})$	8.5	6.5	5.5	6.8
$N_{\text{D}}(\text{cm}^{-3})$	1.6×10^{17}	2.4×10^{17}	1.4×10^{17}	1.6×10^{17}
$V_{\text{BE}}(\text{V})$	39	38	8	6
$f_{\text{T}}(\text{GHz})$	33	29	34	43
$f_{\text{max}}(\text{GHz})$	79	48	40	39
$I_{\text{CE}}(\text{mA})$	20	20	20	23
$h_{\text{FE}}(\text{V})$	2.8	2.8	2.8	3.8

Table 6-2 Device Characteristics for Constant Thermal Co

Device(Ge)	0.15%	0.30%	0-1.5.2%	0-12.2%
$W_{\text{base}}(\mu\text{m})$	26	36	33	48
$N_{\text{A}}(\text{cm}^{-3})$	5×10^{18}	5×10^{18}	5×10^{18}	5×10^{18}
β	94.2	118	128	66
$W_{\text{emitter}}(\mu\text{m})$	0.5	0.5	0.5	0.8
$N_{\text{D}}(\text{cm}^{-3})$	2.6×10^{17}	2.6×10^{17}	2.6×10^{17}	2.6×10^{17}
$V_{\text{BE}}(\text{V})$	11.5	90	60	81
$f_{\text{T}}(\text{GHz})$	78	66	20	36
$f_{\text{max}}(\text{GHz})$	58	55	35	43
$I_{\text{CE}}(\text{mA})$	18	14	15	18
$h_{\text{FE}}(\text{V})$	2.5	3.0	2.4	3.3

We therefore maximize efforts of relaxing the design constraints on N_d and β . Instead, we will maximize the Ge content at the collector side of the base and try to maintain V_A . We will thus keep N_d high, to 10^{18} cm^{-3} as in the last column of Table 6-1, which is almost the upper limit defined by tunneling in the emitter-base junction. This will maximize E_{bc} , as well. Note here that the Ge grading is limited by the stressed layer stability criterion for a given base region width. The first column in Table 6-2 reflects the MMSPICE-3.0 simulations of that modified device, evolved from the last column of Table 6-1. The predicted performance is reasonably good, except for the high β and somewhat low BF_{CEQ} . The current gain must be increased, and with the emitter region fixed, the only way that can be done is to widen W_{BB} . The next three columns in Table 6-2 reflect the simulation results. Note that because of the limit on total Ge, the Ge content at the collector side has to be lowered as W_{BB} is increased. Hence the effective base channel number, which defines β , is increasing because of the relaxed Ge as well as the increased N_d .

The third-column data in Table 6-2 show near-optimal performance for mixed mode applications, i.e., moderate β and V_A for the analog application and modified device speed for ECL circuits, but with slightly low BF_{CEQ} . In fact, the device reflected the column is comparable to the optimal design recently developed at IBM (Finf00), the technology of which has been applied to the fabrication of record-high performance of 12-bit DOL converters. The IBM device has a triangular-Ge profile with the following device characteristics: β of 70, V_A of 7V, peak f_T of 40GHz, peak f_{MAX} of 60GHz, BF_{CEQ} of 3.0%, and maximum ECL propagation delay t_{pd} of 30.7ps. This favorable comparison proves that the MMSPICE-3.0-predicted characteristics are reasonable, and thus supports the utility of our HBT model in IC TCAD. Since our simulations include reasonable parameters associated with the extrinsic device region, the

predicted improvement in RC propagation delay over the 1BM device is noteworthy. From the simulation results in Table 6-2, we can conclude that the minimum/low grading coefficient N_A are desirable for achieving maximum speed and large V_A . If the required speed cannot be achieved by base-region optimization alone, then emitter-region design could be considered at this stage in order to reduce the emitter transit time and perhaps give more flexibility in lowering β .

Flexibility in lowering β and controlling BF_{CEO} might be affected by collector-region design optimization, which we have not yet considered. As discussed in Chapter 4, the high doping density in the *epi*-collector of the NPT device negates any dependence of BF_{CEO} on N_{EP} . Thus to increase BF_{CEO} at all, we must reduce N_{EP} , which lowers the peak electric field at the collector junction and hence tends to reduce the impact ionization (see Chapter 6). The device and circuit performance data in Table 6-2 reflect the importance of MINIMEX-3.1 simulations in this regard. Indeed lowering N_{EP} increases BF_{CEO} . Of course there is a lower limit of acceptable N_{EP} imposed by quantum effects and associated reduction in high-current speed. The minimum $i_{c,s}$ will come at lower currents as N_{EP} is reduced. The required driving current for maximum device speed dictates the minimum allowed N_{EP} . If this N_{EP} does not satisfy the required maximum BF_{CEO} , we could consider the insertion of a low-doped layer between the base and *epi*-collector regions to reduce the peak electric field intensity at the collector junction (Ladd). We note in Table 6-2 that V_A and f_{BCE} also increases significantly with the N_{EP} reduction, due to the right trend seen both of these figures of merit (inversely proportional to the collector-base junction capacitance). Also with this change, the IPW effect tends to diminish because of the more symmetrical collector-base junction.

Table 8-3: Device Characterization with Epi-Collector Doping Level Variation

Device(Dc)	0-15.5% 0-15.5%	0-17.5% 0-17.5%	0-19.5% 0-19.5%	0-21.5% 0-21.5%
$W_{\text{base}}(\mu\text{m})$	50	50	50	50
$N_{\text{A}}(\text{cm}^{-3})$	9×10^{18}	9×10^{18}	9×10^{18}	9×10^{18}
β	128	128	128	128
$W_{\text{epi}}(\mu\text{m})$	0.5	0.5	0.5	0.5
$N_{\text{D}}(\text{cm}^{-3})$	2.5×10^{21}	2.5×10^{21}	2.5×10^{21}	2.5×10^{21}
$V_{\text{BE}}(\text{V})$	98	108	204	1400
$f_{\text{T}}(\text{GHz})$	50	50	46	50
$f_{\text{max}}(\text{GHz})$	50	50	38	45
$I_{\text{B}}(\text{mA})$	15	14	14	15
$h_{\text{FE}}(\text{V})$	3.4	3.8	4.3	5.8

Table 8-4: Device Characterization with Epi-Collector Thickness Variation

Device(Dc)	0-11.5% 0-11.5%	0-15.5% 0-15.5%	0-19.5% 0-19.5%	0-23.5% 0-23.5%
$W_{\text{base}}(\mu\text{m})$	50	50	50	50
$N_{\text{A}}(\text{cm}^{-3})$	9×10^{18}	9×10^{18}	9×10^{18}	9×10^{18}
β	128	128	128	128
$W_{\text{epi}}(\mu\text{m})$	0.2	0.3	0.5	0.7
$N_{\text{D}}(\text{cm}^{-3})$	2.5×10^{21}	2.5×10^{21}	2.5×10^{21}	2.5×10^{21}
$V_{\text{BE}}(\text{V})$	1115	1040	778	241
$f_{\text{T}}(\text{GHz})$	50	50	50	50
$f_{\text{max}}(\text{GHz})$	45	50	45	31
$I_{\text{B}}(\text{mA})$	15	14	13	12
$h_{\text{FE}}(\text{V})$	3.3	3.6	4.1	4.4

We note that the overall column data in Table 4-3 reflects a reasonably optimized technology for both analog and digital applications. The third- and fourth-column data show even better performance in V_A , BF_{CEO} and f_{TAC} . However, W_{BPT} is much thicker than the base and emitter widths, and the defined device is probably expected as highly scaled technology. We then use MINIPICE-3.0 to define the minimum W_{BPT} allowed by the breakdown requirement. The data in Table 4-4 show effects of reducing W_{BPT} in the Table 4-3 device from 0.3 μm to 0.2 μm . Except for the first-column data, all cases show still reasonable characteristics for mixed mode circuit applications, with good breakdown voltages. We consider the fourth-column device as the optimal case as it satisfies all the requirements which we initially intended for the mixed-mode applications, although the other two cases can be considered as optimal depending on the specific needs. We note that the different V_A trend in Table 4-4, in contrast to that in Table 4-3, reflects changed charge distribution in the emitter-collector with W_{BPT} reduction, from the conventional forward-active mode to the contemporary forward-active mode. Also the uniformity in f_T characteristics implies that the emitter and base transit times dominate device speed.

In Fig. 4-2, MINIPICE-3.0 predicted f_T versus I_C plots are illustrated for both the optimized device and the actual NTT device. It is clearly seen that the IBVBI effect is reduced and the current driving capability is increased even with the lower emitter-collector doping. The f_{TMAX} and BCL f_{T0} are also improved substantially as shown in Figs. 4-3 and 4-4, respectively. The Early voltage V_A improvement is also evident as shown in Fig. 4-5. Note the remarkable flat I-V characteristic in Fig. 4-6 compared to that in Fig. 4-3 for the NTT device.

To illustrate the impact of this improvement more clearly at the circuit level, we consider a simple differential-amplifier circuit shown in Fig. 4-8. One of the important figures-of-merit of this circuit is the common-mode rejection

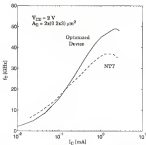


Fig. 4-5 Measured f_p vs i_D characteristics of NTT device and the calculated one of the optimized device. Note the increased driving current level as well as the peak f_p .

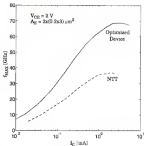


Fig. 6-3 Measured f_{max} vs I_C —characteristics of NTF device and the simulated f_{max} of the optimised device. Note the increased driving current level as well as the peak f_{max} .

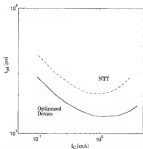


Fig. 4-4. DCL propagation delay measured with NTT device and the optimized device.

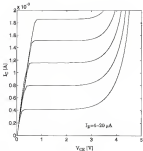


Fig. 4-5 Output characteristics of the optimized device. Note the extremely flat $I-V$ characteristics, which implies very large Early voltage.

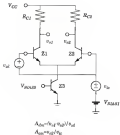


Fig. 4-6 A simple differential amplifier circuit. Differential-mode input signal v_{id} , v_{od} and common-mode input signal v_{ic} are schematically shown. Also differential-mode gain A_{dm} and common-mode gain A_{cm} are defined.

ratio (CMRR), which is defined as the ratio of the differential gain A_{dm} to the common-mode gain A_{cm} , that is,

$$\text{CMRR} = \frac{A_{dm}}{A_{cm}} = \frac{(v_{o,d} - v_{o,c})/v_{i,d}}{v_{o,c}/v_{i,c}} \quad (4-12)$$

where $v_{o,d}$ and $v_{o,c}$ are outputs taken from the collector terminals of the differential stage, $v_{i,d}$ and $v_{i,c}$ are differential-mode input and common-mode input, respectively. Ideally, a differential amplifier is expected to be insensitive to common-mode (i.e., symmetrical) inputs, and amplify only the differential mode (i.e., antisymmetrical) inputs. Thus the higher the CMRR, the closer the circuit is to being an ideal differential amplifier, which only amplifies differential-mode signals. The CMRR of this circuit is approximately equal to $V_{A,25}/V_T$ at low frequencies, where $V_{A,25}$ is the Early voltage of the device 25 in Fig. 4-4, and V_T is the thermal voltage. As shown in Fig. 4-7, the predicted CMRR of the circuit based on our optimized device is substantially larger than that predicted for the HBT technology at low frequencies. Actually, V_A is a representative scaling figure of merit only for relatively low frequencies. At high frequencies, the small-signal impedance seen through the collector terminal of the device 25 (Fig. 4-4) is in parallel with the small-signal output resistance represented by V_{A2}/I_C . Thus the effective output impedance of the device is decreased at high frequencies, which accounts for the degradation of CMRR in Fig. 4-7. Nevertheless, typical operating frequencies are relatively low, and the performance improvement implied for the optimized device in Fig. 4-7 is significant.

Through this simulation study, we showed that SiGe HBT technology can provide superior performance for the mixed-mode circuit applications by engineering base and collector structures properly. We demonstrated that our new SiGe HBT model can be useful as a design tool for technology CAD.

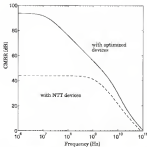


Fig. 6-7 Frequency-dependent characteristic of Common Mode Rejection Ratio of the differential amplifier circuit in Fig. 6-6, was added with MTT device and with optimized device.

Combined with the device simulation capability, this model can give quick insight on the device design from the device level, which is highly desirable in customizing technology development.

6.6 Summary

We described an application of SiGe HBT model to design a high performance device for mixed mode applications. Beginning with the MTT technology, we modified the base region first to improve the device speed in terms of f_T and the Early voltage V_A , which are both important in analog applications. The maximum performance of these two was obtained by allowing maximum base region width and maximum Ge grading within the limit of the strained layer stability. In determining the optimal base region width, we considered the current gain and the emitter-base junction tunneling, which limits the peak base doping density. Then we proceeded to the collector design to improve the breakdown BV_{CEO} . We also discussed qualitatively the variation in f_{MAX} and V_A with the change in the np-collector doping level. Exemplary device characteristics were given to illustrate the trade-off involving various structural parameters. Through the design example, we demonstrated that our model is very useful as a TCAD tool in developing new SiGe HBT technology.

CHAPTER 7 SUMMARY AND SUGGESTIONS FOR FUTURE STUDY

7.1 Summary

This dissertation addressed the modeling, simulation, and design of the SiGe-based HBT for applications in high-performance integrated circuits. The significant achievements made through this work can be summarized as follows.

First, with an objective of enhancing MMSPICE to eliminate its drawbacks and to make it a basis for the SiGe HBT model, we optimized the regional semi-empirical device model formulation with regard to accuracy, convergence, and computational efficiency. The details of the model system modifications implemented in MMSPICE-4.0 were presented with device and circuit simulation results to demonstrate the significant improvement in the simulator.

Second, we developed a physical but compact model for the SiGe-based HBT which accounts for the nonuniformities in base doping profile and Ge grading and is applicable to the conventional HBT. Also the energy barrier formation at the collector heterojunction was modeled physically. The developed model was implemented into SPICE3.0 for circuit simulation, creating MMSPICE-4.0. New model parameter evaluation schemes were developed and described.

Third, a semi-local accounting of impact-ionization current was developed based on a simplified energy-balance equation and the electric field

obtained as a by-product of transport analysis in the model routine. The electron temperature was calculated analytically and the multiplication factor was calculated by numerically integrating the ionization rate across the collector region, including buried layer. The model was implemented as post-processing to make it computationally efficient. Through the predictive results based on the existing technologies, we showed breakdown behavior can be sensitive to the epi-collector thickness depending on the epi-collector doping level.

Fourth, to assess the model implementation, we applied our model to a highly scaled aggressive 11T SiGe-base HBT technology. Simulated results with and without the Ge incorporated were given to show that our model reliably accounts for the bandgap narrowing effect due to the Ge as well as the heavy doping. Through the measured and simulated dc characteristics, we found that the peak current where the maximum device speed is achieved is significantly lower than the (theoretical) limit for the base pushout, and inverse base-width modulation (IBWM) – a widening of the quasi-neutral base width – and collector SCR-width widening (SCRW) in highly scaled 11Tn and 11Tb were observed and analyzed as important mechanisms governing scaled device speed degradation at high currents.

Finally, to demonstrate the utility of the SiGe HBT model as a TCAD tool in optimizing device and circuit performances, we used it to explore the performance limit of this technology by varying structural parameters, and compared the results with published data. With a systematic design procedure, we achieved reasonable figures of device and circuit performances for mixed mode applications.

1.2.8 Suggestions for Future Study

First, the present HBT model assumes a conventional emitter structure, which has higher doping than the base region. To reduce the tunneling current at the emitter-base junction that can occur at high base doping levels, an intrinsic region could be inserted between the emitter and base regions, forming a high-low emitter doping profile. The proper emitter-region model to construct for this is needed to extend the model validity further.

Second, IED research groups [Kucuk, Ozturk] report experimental results of the Ge-based HBTs with the Ge profile linearly graded in part and flat for the rest of the base region. In spite of the adequacy of the present model, which accounts for trapezoidal (including triangular) Ge profiles as well as a flat one, accounting for the above mentioned Ge profile might be included in the model to extend its applicability.

Third, the potential barrier formation model requires a refinement and accordingly refinement. There seems to be a convergence problem related to this model mainly due to the flat collector-current characteristics that can result in the Gummel plots. The physical significance of the barrier in device and circuit simulation also needs to be studied further.

Fourth, the present model cannot guarantee the continuity of derivative of the terminal current across model boundaries, which sometimes results in abrupt change in its characteristics such as I_C - I_E plots. The main source of this abnormality seems to be related to the modeling of the quasi-Fermi potential separation, V_{BC} , at the collector-side moving boundary of the QNS region. Thus, it is suggested to refine this model more physically to solve this problem.

Finally, the present model equates the actual base doping profile to an exponential profile, which might make the junction QNS modulation behavior

less accurate, especially at the collector junction. Although the present model is good enough for the device with uniform base doping, further refinement may be required to better describe unique phenomena in the SiGe HBT such as Early and reverse-Early effects [Petrík, Grutka]

APPENDIX A EFFECTIVE BULKEN FIELDS DEFINED BY BANDGAP NARROWING IN CASE REGION

With negligible hole current, the hole-current equation yields the built-in field for holes

$$E_p = \frac{D_p}{\mu_p \rho} \frac{d\rho}{dx} = \frac{1}{q} \frac{dE_F}{dx} \quad (\text{A-1})$$

Then, the built-in field acting on the electrons can be expressed in terms of the bandgap parameters or effective carrier density as follows

$$\begin{aligned} E_n &= \frac{1}{q} \frac{dE_C}{dx} = \frac{1}{q} \frac{d(E_V + E_g - AE_g)}{dx} \\ &= E_p - \frac{1}{q} \frac{dAE_g}{dx} \\ &= \frac{D_p}{\mu_p \rho} \frac{d\rho}{dx} - \frac{1}{q} \frac{dAE_g}{dx} \\ &= \frac{kT}{q} \left[\frac{1}{\rho} \frac{d\rho}{dx} - \frac{1}{kT} \frac{dAE_g}{dx} \right] \quad \text{Einstein relation under non-degeneracy} \\ &= \frac{kT}{q} \left[\mu \exp\left(-\frac{AE_g}{kT}\right) \right]^{-1} \exp\left(-\frac{AE_g}{kT}\right) \left[\frac{d\rho}{dx} - \frac{\rho}{kT} \frac{dAE_g}{dx} \right] \\ &= \frac{kT}{q} \frac{d}{dx} \ln \left[\rho \exp\left(-\frac{AE_g}{kT}\right) \right] \\ &= \frac{kT}{q} \frac{d}{dx} \ln \left[\rho \frac{n_i^2}{n_{in}} \right] \\ &= \frac{kT}{q} \frac{d}{dx} \ln [\rho_{eff}] \quad \text{with } \rho_{eff} = \rho (n_i^2/n_{in}^2) \end{aligned} \quad (\text{A-2})$$

APPENDIX B ELECTRON TEMPERATURE DISTRIBUTION IN EPI-COLLECTOR

In general, the epi-collector comprises two different regions characterized by their electric field distributions as indicated in Fig. 9-1. In each region $E(x)$ can be approximated as linear, including $E(x) = 0$ in the buried layer. Hence $E(x)$ can be described in a general format as follows:

$$E(x) = Ax + B \quad 0 \leq x \leq W \quad (B-1)$$

$$= 0 \quad (x - W) \leq D \quad W < x \leq W_{\text{epi}} \quad (B-2)$$

$$= 0 \quad x > W_{\text{epi}} \quad (B-3)$$

where $x = 0$ is at the collector base metallurgical junction, and $x = W$ divides the epi-collector region. The coefficients A through D and W for each mode depend on I_C and V_{CB} as given below:

I) Classical Forward Active Mode ($I_C < I_{\text{BCR}}$ and $V_{BCR} < V_{BC} < V_{\text{CCR}}$, or $I_C < I_{\text{CCR}}$ and $V_{\text{CCR}} < V_{BC} < V_{\text{CCR}}$)

$$A = (1/Am_0)(I_C - I_0), \quad B = E_0 - AW_{\text{BCR}}, \quad C = 0, \quad D = E_{\text{CCRBCR}}, \quad \text{and} \\ W = W_{\text{BCR}}$$

II) Ohmic Quasi-Saturation Mode ($I_{\text{CCR}} < I_C < I_0$ and $V_{\text{CCR}} < V_{BC} < V_{\text{CCR}}$)

$$A = B = C = 0, \quad D = E_{\text{CCRBCR}}, \quad \text{and} \quad W = W_{\text{BCR}} - W_{\text{CCRBCR}}$$

III) Non-Ohmic Quasi-Saturation Mode ($I_0 > I_{\text{CCR}}$ and $V_{BC} < V_{\text{CCR}}$)

$$A = B = 0, C = (1/\Delta\omega_g)(I_C - I_Q), D = E_Q, \text{ and } F = W_{JEF} - W_{JCF}$$

4) *Contemporary Forward-biased Mode* ($I_{JCCF} = I_C + I_{JCCF}$ and

$$V_{JCF} < V_{JCCF})$$

$$A = B = 0, C = (1/\Delta\omega_g)(I_C - I_Q),$$

$$D = (I_Q - V_{JCF})/(W_{JCF} - W_{JEF})(I_C - I_Q)/\Delta\omega_g, \text{ and } F = 0$$

The critical currents and voltages used above in defining each operational mode are defined as follows:

$$V_{JCCF} = I_Q - I_Q R_{JEF},$$

$$I_{JCCF} = I_C + \Delta\omega_g(V_{JCF} - V_{JCCF})/W_{JCF},$$

$$I_{JCF} = I_Q + \Delta\omega_g(V_{JCCF} - V_{JCF})/W_{JCF},$$

$$I_{JCF} = I_Q + (V_{JCCF} - V_{JCF})/R_{JEF}$$

Since $\mathcal{E}(x)$ is linear in each region, analytic expressions for $T_a(x)$ defined by (4-5) can be derived, they are given in the general format as follows

$$\frac{\partial \mathcal{E}}{\partial x} \bigg|_0 (T_a(x) - T_b) = Ax - (Ak_E - B) \left(1 - \exp\left(-\frac{x}{\lambda_g}\right)\right), \quad 0 < x < W \quad (B-6)$$

$$\begin{aligned} \frac{\partial \mathcal{E}}{\partial x} \bigg|_0 (T_a(x) - T_b) &= \exp\left(-\frac{x}{\lambda_g}\right) (F^2 + G(x - W)) \\ &\quad - (C\lambda_g - D) \left(1 - \exp\left(-\frac{x - W}{\lambda_g}\right)\right), \quad W < x < W_{JCF} \end{aligned} \quad (B-7)$$

$$\frac{\partial \mathcal{E}}{\partial x} \bigg|_0 (T_a(x) - T_b) = \exp\left(-\frac{x}{\lambda_g}\right) (F^2 + F\lambda), \quad x > W_{JCF} \quad (B-8)$$

where

$$F_1 = \exp\left(\frac{W}{k_B}\right) \left[(\Delta W + (\Delta k_B - D) + 1) \exp\left(-\frac{W}{k_B}\right) + 1 \right] \quad (2.17)$$

and

$$F_2 = C(W_{\text{eff}} - W) \exp\left(-\frac{W}{k_B}\right) + (2k_B - D) \left[\exp\left(-\frac{W}{k_B}\right) + \exp\left(\frac{W}{k_B}\right) \right] \quad (2.18)$$

APPENDIX C CURRENT GAIN AND EARLY VOLTAGE IN SiGe HBT

The dependence of collector current of SiGe HBT on collector-base bias can be formulated in terms of the effective base Gummel number using the new integral equations for a drive base (KroEM)

$$I_C = qA_{\text{base}} \exp(qV_{BE}/kT) \left[\int_0^{W_B} \frac{p}{D_B n_{iB}^2} dx \right]^{-1} \quad (\text{C-1})$$

where the integration is performed over W_B , the width of quasi-neutral base region width.

Variations in the base width with voltage V_{CB} causes a variation in the collector current that can be written

$$\frac{\partial I_C}{\partial V_{CB}} = -qA_{\text{base}} \exp(qV_{BE}/kT) \frac{p(W_B)}{D_B (W_B) n_{iB}^2 (W_B)} \left[\int_0^{W_B} \frac{p}{D_B n_{iB}^2} dx \right]^{-2} \frac{dW_B}{dV_{CB}} \quad (\text{C-2})$$

which represents the small signal conductance of the collector-base junction and also can be written in terms of I_C

$$\frac{\partial I_C}{\partial V_{CB}} = -I_C \frac{p(W_B)}{D_B (W_B) n_{iB}^2 (W_B)} \left[\int_0^{W_B} \frac{p}{D_B n_{iB}^2} dx \right]^{-1} \frac{dW_B}{dV_{CB}} = -\frac{I_C}{V_A} \quad (\text{C-3})$$

The Early voltage, which is defined by the reciprocal term multiplying the current in (C-3), is

$$V_A = \left[\frac{p(W_B)}{D_B (W_B) n_{iB}^2 (W_B)} \frac{dW_B}{dV_{CB}} \right]^{-1} \int_0^{W_B} \frac{p}{D_B n_{iB}^2} dx \quad (\text{C-4})$$

Since the total base majority charge Q_B across the quasi-neutral base in the active base is written

$$Q_B = qA \int_0^{W_B} n_0 dx, \quad (C-6)$$

the derivative of (C-5) with respect to V_{CE} is

$$\frac{\partial Q_B}{\partial V_{CE}} = \Delta C_{CB} = qAq_n(W_B) \frac{\partial W_B}{\partial V_{CE}} \quad (C-7)$$

where C_{CB} represents the small signal collector-base junction capacitance per unit area. Substituting (C-6) into (C-6), the Early voltage V_A can be written

$$V_A = \frac{qQ_B(W_B)^2 n_0^2(W_B)}{C_{CB}} \int_0^{W_B} \frac{p}{D_n n_0^2} dx \quad (C-8)$$

Assuming the back injection current does not change significantly with the SiGe base layer characteristics, the current gain β is proportional to the collector current and is given by

$$\beta = qA n_0^2 \left[\int_0^{W_B} \frac{p}{D_n n_0^2} dx \right]^{-1} \quad (C-9)$$

where $qA n_0^2$ is the base saturation current. Multiplying (C-8) and (C-9), an important relation between the current gain and Early voltage of a SiGe HBT is obtained [Fackl]

$$\beta V_A = \frac{q^2 D_n(W_B) n_0^2(W_B)}{J_{BE} C_{CB}} = \frac{q^2 D_n(W_B) n_0^2}{J_{BE} C_{CB}} \exp\left(\frac{qV_A(W_B)}{kT}\right) \quad (C-10)$$

where $kV_A(W_B) = q_n^2(W_B)/n_0^2$ is the total apparent bandgap narrowing given as the sum of each contribution by heavy doping and Ge.

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